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linear amplifier considerations

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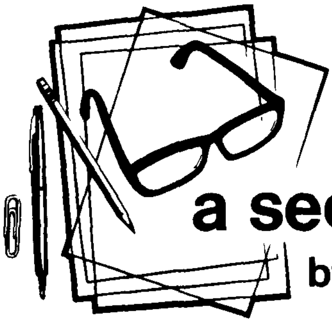


a second look

by Jim Fisk

Beginning with a small handful of amateur wireless operators in the early 1900s, the Amateur Radio population in the United States has grown to the point where it's now approaching 360,000. Last year the growth rate was about 8 per cent, down slightly from 1977, and this year it is expected to be about the same. And while *orderly* growth is healthy for the hobby, in many ways the Amateur Radio Service is like the proverbial "house that Jack built," with rooms added as they are required, with little thought to future construction — or indeed, to the esthetics of the architecture!

If you study the history of Amateur Radio, it's easy to understand why this happened: for years there were more licensed Amateur Radio stations in this country than all the other radio services combined, many top members of the FCC were hams, and the management ranks of most major radio-electronics firms were filled with licensed amateurs — many, in fact, began their careers as Radio Amateurs. With influential friends in high places who had a vested interest in Amateur Radio, most operators gave little thought to the future. The complexion of Amateur Radio has changed over the years, however, and it's obvious that we can no longer afford such a *laissez faire* attitude toward our future.

One matter that concerns many of the older hams is that in the past 25 years the character of Amateur Radio has evolved slowly away from being a technician's hobby, where much of the operating equipment was homebuilt, to an operator's hobby, where little or no technical expertise is required. This is not necessarily a problem because our activities are closely linked not only to a rapidly changing technology, but to a dynamic society that continually confronts Amateur Radio with new obstacles, challenges, and opportunities for providing useful public service. Nevertheless, more thought must be given to the impact of this trend on the long range future of Amateur Radio.

With a steadily increasing number of amateurs and greater government intervention in terms of changed licensing regulations, restrictive antenna covenants, and RFI requirements (not to mention WARC 79 and the proposed revision of the Communications Act), it's increasingly apparent that *all* of us must give some serious thought to where the Amateur Radio Service should be in the coming decade. While long-range planning is hardly an exact science, it is possible to anticipate some of the problems, to perceive certain distant opportunities, and to develop appropriate recommendations. If we put our collective heads together, we should be able to plot a positive future course for Amateur Radio — rather than drifting out of control as we have for the past few years, reacting to external events as they have occurred. Positive results, however, will require a substantial amount of effort on a continuing basis by a large number of concerned amateurs. Complaining about the current state of affairs or railing about the "system" in the press is neither positive nor constructive.

Those of you who have read my editorials for the past eleven years know that I have pointedly avoided the politics of Amateur Radio. Therefore, when I suggest that a possible focus for future planning activities is the ARRL's Long-Range Planning Committee, you know that suggestion is not politically motivated. For those of you who are not members of the ARRL, the Long-Range Planning Committee was established by the ARRL Directors in January for the purpose of "reviewing and making recommendations concerning programs which the ARRL is and should be providing to its members and to the Amateur Radio Service . . ."

At its initial meeting in February the members of the committee, according to one of those present, agreed upon several criteria which would govern the committee's activities:

1. The general welfare of the entire Amateur Radio Service was to be served, not just parts of it.

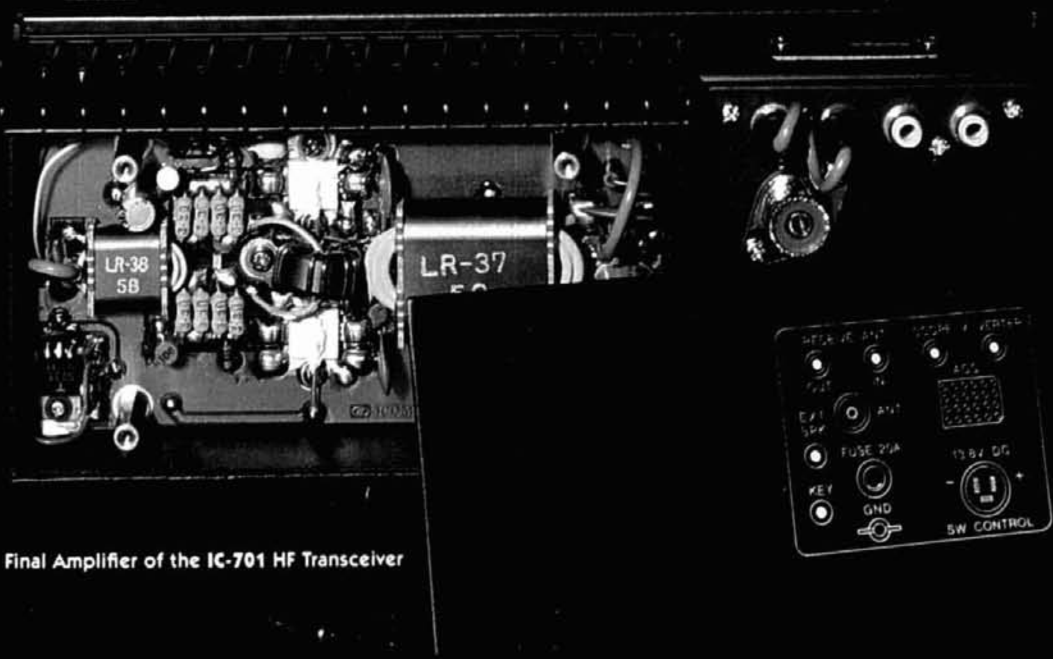
2. No fact of the ARRL's operation was exempt from scrutiny.

3. A subject as complex and far reaching as the future of Amateur Radio cannot be properly appraised without inputs from many different people — *ARRL members or not*.

If *you* have any comments or recommendations about the future of Amateur Radio, make it a point to let the Long Range Planning Committee (LRPC) have the benefit of your thoughts. A letter or card to Vic Clark, W4KFC (12927 Popes Head Road, Clifton, Virginia 22024), marked for the attention of the LRPC, will be acknowledged, and Vic will make sure that your comments are available to each of the members of the committee.

Jim Fisk, W1HR
editor-in-chief

The Final Warranty



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comments

longhand printed-circuit layout

Dear HR:

The article, "Printed-Circuit Layout Using the Longhand Method," which appeared in the November, 1978, issue of *ham radio* was especially interesting to me because I have been using the described technique for the past two years. However, the additional steps which I use will produce a better board and are desirable with high-density boards. First, lines on the paper pattern which represent the copper should be made red and the circuit components should be shown black. This makes reading and interpretation easier. Second, when the pattern is transferred to the copper foil surface by marking with a sharp point, the points should not be deep enough to produce burrs (burrs tear the pen point and eventually make it difficult to produce a clean fine line). Third, after the points are on the copper those which are connected should be joined by a pencil line. This is important on a high-density board because it permits inking the lines rapidly and the pen point does not have time to dry. Lines produced with a dry point must be retraced; a smeared line is nearly always the result.

Maximum ground-plane area is a requirement for rf circuits, especially for vhf circuits. Filling in all the blank area with a pen is tedious and it's difficult to make the area completely resistant to the etch solution. Tape can be used but covering small irregular areas is difficult. I find that an easy method is to outline the ground plane area first with the pen, then fill in all the enclosed area with the blue dye used by tool makers and machinists. I have tried many inks and paints for this step, and the blue machinists dye is superior to all others. It is easy to apply with a small brush and dries rapidly. The board may be etched minutes after application. A coat which appears too thin will resist the etchant even at elevated temperature. The ink wets the copper surface and flows easily but stops when it contacts the previously applied line.

The machinists dye is also useful for producing plug patterns on circuit boards. Coat a 1-cm (½-inch) strip at the board edge and use a sharp point or jeweler's screwdriver to remove the ink from areas between the contacts. Use an old plug, placed against the board edge, for a pattern. It is easy to produce 22-contact plug patterns with this method.

I have used two types of the blue dye. One is called *Dykem Steel Blue* and is a product of Dyken Corporation of St. Louis, Missouri. Another is called *Mike-O-Blue* and is sold by Ashburn Industries of Houston, Texas. A four-ounce can is adequate for many boards.

This "longhand" method will produce high-quality circuit boards which have clean lines and ground-plane areas without pit marks commonly found on boards which have been prepared by other methods. Boards with 3-mil copper foil can be etched in less than ten minutes in a 50 per cent etch solution (ferric chloride) heated by placing it in a tray or plastic dish floated in hot water. Use only enough etch solution to cover the board about 1 cm (½ inch) and agitate during etching to provide a washing action. In addition, the copper surface can be seen during the process so the board may be removed when completed.

I have tried many types of ink pens and find that the *Sharpie* brand is best; their number 49 has the best point. Store the pen with the point down, this aids in keeping a generous supply of ink in the point and is always ready to use.

During the past years I have spent many hours trying to find an easy method for producing "longhand" etched circuit boards and have concluded that the technique described in *ham radio*, along with the additions indicated above, is the best.

Robert J. Grabowski, W5TKP
Houston, Texas 77005

Dear HR:

During a literature search for an electronics project I recently went through my file of *ham radio* — and was distracted for three evenings reading the fine articles in three years of issues! Yours is by far the highest quality journal of all those devoted to Amateur Radio; please don't compromise that quality.

Guy Rothwell, KH6JCD
Kailua, Hawaii

Dear HR:

I just breadboarded the CW processor described by Jones in the October, 1978, issue of *ham radio*. It's really sharp! However, the 555 oscillator interacts with the operation of the 576. To cure this problem, I've installed a 100-ohm resistor between the 5-volt line and pin 8 of the 555, and a 220-μF capacitor between pin 8 and ground. It makes a real improvement.

Jeff Davis, VE3CBJ
Grimsby, Ontario

presstop

ST. VINCENT'S VOLCANIC ERUPTION in the middle of April essentially destroyed conventional communications throughout the Caribbean island as it left over 15,000 homeless. Amateur Radio immediately filled the breach, beginning when VP2s SQ, SHE, SAZ, SK, and others set up a communications center at police headquarters. Other Amateurs operating from police stations and relief centers throughout the island have since provided almost all communications for the hard hit nation.

Island Premier Milton Cato relayed a message to Miami Coast Guard via KV4FZ requesting help which was immediately forthcoming with three helicopter-equipped Coast Guard cutters, the Gallatin, Vigilent, and Dallas. Herb also requested an emergency OK for third-party traffic from the FCC and State Department — it was forthcoming in only 1½ hours! *Following the OK*, 7152 (daytime), 3802 (night), 14287 (health and welfare via KA2CPA) and 14313 were all busy with emergency traffic.

Coast Guard And Military, along with the Amateurs, all used the Amateur bands for essential communications, K5OPG/MM on the Gallatin, while the other cutters (without licensed Amateurs) were using their Coast Guard identification. C-130 transports from the Canal Zone, bringing in relief supplies, communicated principally on 40 meters.

Amateur Bands also supplied inter-island communications. 8P6AA and 8P6AH have a link operating between the embassy in Barbados and St. Vincent police headquarters, handling all the relief traffic with that near-by country.

RECENT 27-MHZ EXPANSION PROPOSALS could actually threaten the Amateur 10-meter band as the Washington scene appears to be shaping up. Both the CB Magazine proposal for adding SSB CB channels (RM-3299) and the Washington State CB Radio Association's Petition for a new hobby class "Amateur" allocation below 28 MHz (RM-3317) are reported receiving very heavy positive response, almost entirely from CBers, CB organizations and (assumedly) those already active in the 27405-28000 kHz spectrum. The comments, almost without exception, endorse the merits of more flexible, higher power, longer range communications, and often cite a basic, no-code Amateur license as part of the package.

The Number Of Respondents and their enthusiasm seem to have caught the Commission's attention, from staff level up. Japan's experience, where very elementary Amateur licenses have spawned an Amateur population about twice that of the United States, and the experiences of some other nations such as Russia which also have entry level licenses are also being looked at closely.

The Real Threat to the present integrity of 10 meters thus lies in the Washington State petition. Under present international allocations, the frequencies they want for a new "Amateur" service are not Amateur, and it's inconceivable that the FCC would opt to make such an allocation on the eve of a World Administrative Radio Conference. (CB Magazine's proposed CB expansion would, on the other hand, be legitimate now.) The problem with the Washington State proposal is that it could, without too much stretching, lead into a domestic restructuring of 10-meter Amateur use with a rapid influx of new "Amateurs" from 27-MHz SSB ranks.

Nothing Of This Nature is likely to happen in the immediate future due to the imminence of WARC and the current state of the FCC — still in the throes of reorganization. What does seem likely is a new and very hard look at the domestic Amateur service after WARC from a new "consumer oriented" FCC with little knowledge of or concern for traditional Amateur Radio values.

BOTH RUSSIAN SATELLITES may be off the air for good. UA3CR is reported to have told G3IOR in an April conversation — "Both transponders will not be on again!" Telemetry on RS-1's weak beacon has been indicating battery trouble.

Simultaneous Mode A And Mode J operation will be offered by OSCAR 8 on Tuesdays and Fridays, effective June 1. Wednesdays will continue to be Experiment Days, primarily Mode J.

PETITIONS TO RECONSIDER MARITIME mobile's proposed 220 MHz allocation in the U.S. WARC position have been denied by the FCC. The petitions, filed by the 220 MHz Spectrum Management Association of Southern California and seven individuals, were denied by the Commission on the grounds that the WARC position was the result of four years of extensive consideration of the needs and problems of all the services concerned.

Amateurs Received Hard Knocks in Reply Comments filed by the National Radio Astronomy Observatory and the Naval Research Laboratory on the proposed Virginia-West Virginia "quiet zone" (FCC Docket 78-352). In their very substantial filing the government representatives strongly took the League to task for leading the Amateur community astray into a "party line" response. Words like "propaganda," "fiction," and "parrot-like" abound in the government's lengthy rebuttal. While acknowledging Amateur Radio has some value, the generally critical document was almost contemptuous in its dismissal of the efforts of those concerned Amateurs who made the effort to make their feelings known.

THE LONG AWAITED U.S.-HAITI reciprocal agreement was signed April 2 by Haiti, and now awaits State Department action.

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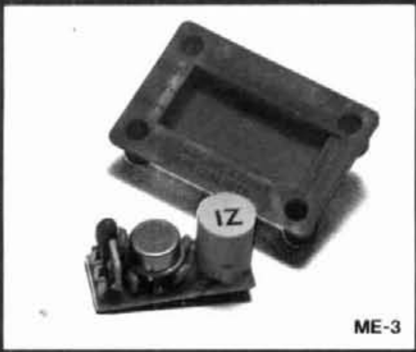

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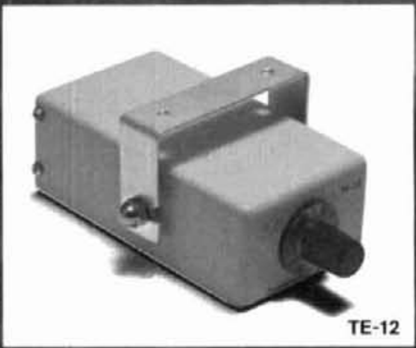

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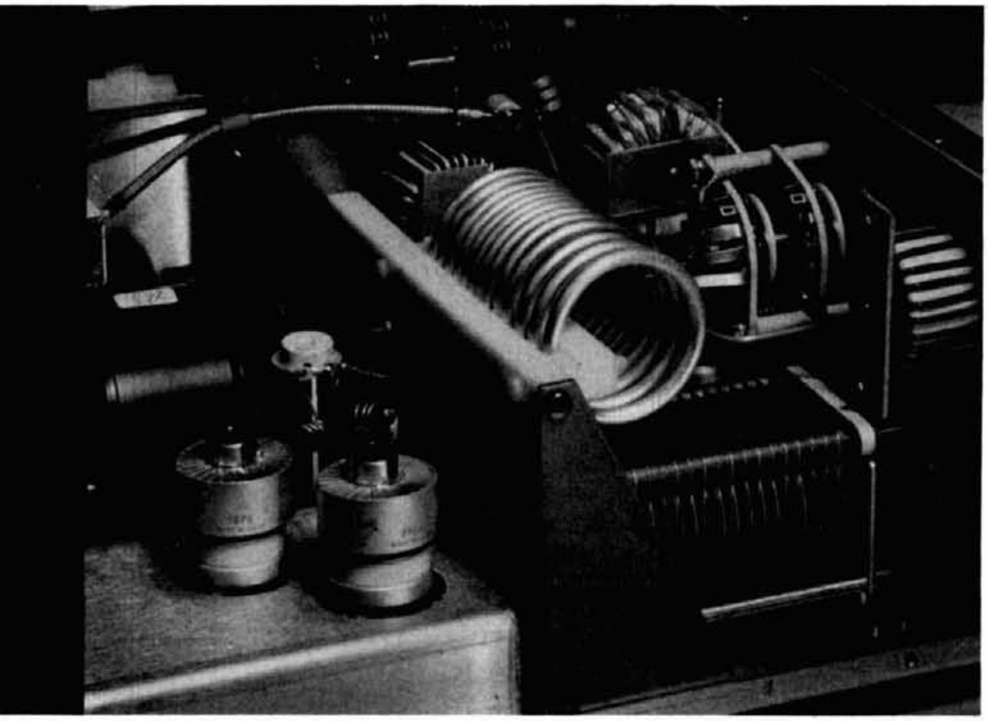


design considerations for linear amplifiers

The first of several articles on practical construction techniques for hf power amplifiers

So you want to build a linear amplifier! So do many other Radio Amateurs. Without doubt, the most popular piece of home-built transmitting equipment (aside from small circuit board projects) is the high frequency linear amplifier. It can be built without having an advanced degree in solid-state technology and computer analysis.

Recent FCC decisions, moreover, have made a linear amplifier homebrew project more inviting to Amateurs, particularly those interested in 10-meter operation. For a period of time, in the early spring of 1978, it was nearly impossible to buy an off-the-shelf, commercial linear amplifier; most manufacturers had stopped production in view of the drastic redesign requirements imposed by the new FCC rules. Home-constructed amplifiers, happily, are exempt from the FCC straitjacket. And, more and more, Amateurs are discovering the fun of building their own amplifiers. It's not as hard as you might think! There's still fun in building and adjusting equipment, and the high-frequency linear amplifier

By William I. Orr, W6SAI, 48 Campbell Lane, Menlo Park, California 94025

described in this series of articles is a good project for the home builder, whether you're an old timer or newly licensed Amateur.

what to build

The builder of a linear amplifier undoubtedly has many questions that must be answerd before he can pick up a soldering iron or drill; what tubes to use, what plate voltage, drive level, harmonic suppression, TVI prevention, cooling, and packaging?

At this stage of the game, even the stoutest of hearts may falter. But cheer up, the overall problem is not complex if the project is approached on a workmanlike basis. The purpose of these articles is to provide a blueprint that will guide the builder through the design, construction, and checkout of a modern linear amplifier capable of operating on all Amateur frequencies between 3.5 MHz and 29.7 MHz. This first article covers design, selection of tubes and components, formulas that make the job easier, and practical construction considerations. A later article will cover the metal-work, assembly, and testing in detail.

Before you jump into the project, bending metal and soldering wires, you should know that there exists a vast amount of literature covering the design and construction of linear amplifiers. It would be foolish to ignore this storehouse of accumulated knowledge. At the end of this article is a list of suggested reading material, and you can learn a lot by observing what has happened in this interesting field of radio design. Since this series of articles cannot possibly cover every detail of building a linear amplifier, you can pick up a lot of very useful extra information if you scan some of the suggested reading material.

preliminary design

The first choice you will have to make concerns the tube (or tubes) to be used, the operating voltages, and the means of cooling the tubes so that their operating temperature will remain within the limits imposed by the manufacturer.

A word of warning is advisable on the subject of surplus or second-hand transmitting tubes. Large power tubes have a finite shelf life. The perfect vacuum has not yet been created, and old tubes (surplus World War II vintage in particular) are not to be trusted — they may have an imperfect vacuum. Surplus tubes marked JAN (which stand for Joint-Army-Navy procurement), such as JAN-813 or JAN-211, provide no warranty to the user, since the tubes are purchased by the military on a special contract with

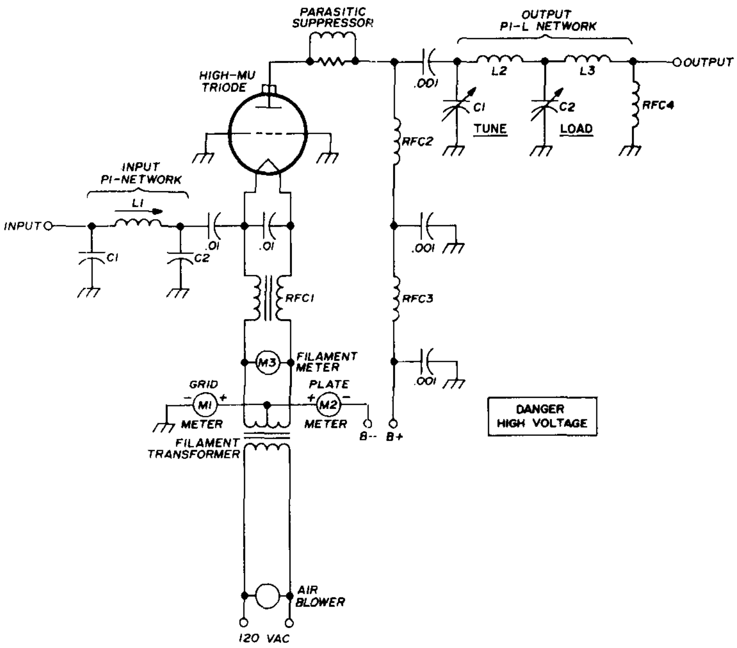


fig. 1. Schematic diagram of a basic grounded-grid amplifier circuit. A high-mu triode tube is used with the exciting signal applied to the filament circuit which has been isolated from the filament transformer and metering circuits by the rf choke, RFC1. A fixed-tuned pi-network circuit matches the output impedance of the exciter to the input impedance of the amplifier. A pi-L plate output circuit is used for maximum harmonic suppression, with a simple parasitic suppressor placed in the plate lead to dampen vhf oscillations. For safety, the metering is placed in the filament return circuit. The grid meter is inserted between the grid (ground) and the filament return, while the plate meter is in the B minus return lead to the power supply. The air blower is connected to the primary of the filament transformer. This circuit may be modified for two parallel-connected tubes by the addition of a second plate parasitic suppressor and increased air blower and filament transformer current capacity. Plate tank circuit components need not be modified if the two tubes run at the same voltage and current as one tube.

source inspection and no warranty return program. Thus, an Amateur who buys a JAN-labelled tube receives no warranty. New tubes purchased from a franchised dealer carry the manufacturer's full warranty. Dealers in surplus tubes, moreover, have no reliable facilities for testing transmitting tubes, which require a large, expensive, and exotic test console. Thus, the purchase of a surplus or second-hand tube may turn out to be penny wise and pound foolish.

To determine the tube type to be used, it is important to note that the most popular ham-type linear amplifiers seen in the various station descriptions and advertisements are capable of running 1-kW input on CW and 2-kW PEP input on ssb. The amplifiers use cathode-driven (grounded-grid) circuitry requiring a drive signal compatible with today's modern exciter (about 80 to 100 watts PEP output). A representative amplifier circuit is shown in **fig. 1**.

Further investigation shows that the modern amplifier concept employs low-profile styling and is designed for desk-top operation next to the exciter. In some cases, the power supply is an external unit. In any case, the amplifier is capable of being controlled by the external VOX or push-to-talk circuit of the exciter. And it can be operated either from 120- or 240-volt primary service.

table 1. Typical rf linear amplifier service, cathode-driven (grounded-grid)

tube	plate voltage	zero signal plate current	maximum signal plate current	maximum signal grid current	maximum signal drive power	typical power output (PEP)
(2) 3-500Z	2500	260	800	240	80	1200
	3000	320	667	230	60	1250
3-1000Z	3000	240	670	220	65	1250
4-1000A	3000	90	670	270	125	1300
8877	3000	130	667	55	50	1150

Representative operating charactersitics of popular tubes suited for cathode-driven service. The 4-1000A is operated as a class-B triode with grid and screen tied together. It can be seen that in terms of efficiency there's not much difference between tube types. The 4-1000A requires the most drive power, the 8877 the least. The differences in power output are insignificant and are within error of measurement. Power output is a function of plate circuit loading and grid drive causing the values to be approximate.

power capability

Given these general specifications, the next step is to determine what goes into the *black box* that is to become your new linear amplifier. If the linear amplifier is to run at 1-kW in the CW mode and 2-kW PEP in ssb service, the choice of tubes to be used narrows. And since cathode-driven (grounded-grid) service is contemplated, selection is restricted to a few tubes which have the ability to sustain this power level with good linearity and low intermodulation distortion. Linear operation implies that the output signal is an exact replica of the input signal; low intermodulation distortion means that unwanted, spurious distortion signals are not generated in or near the signal frequency. When both of these criteria are met, the ssb signal is clean and no power is lost in *furry* sidebands or splatter.

Table 1 shows some practical tubes for linear service and their operating characteristics. From an engineering point of view, there's not much choice between using a single large tube or two smaller tubes in parallel in the high-frequency region. Multiple tubes are thought to be less efficient and generate more intermodulation distortion than one tube, but this is not the case. Power tubes designed for ssb service do not have to be matched pairs, as do the inexpensive TV sweep tubes used in some linear amplifiers.

Regardless of the tube or tubes chosen, grounded-grid triode operation implies class-B service (which you can find outlined in detail in the suggested literature). An important characteristic of this class of service is that tube efficiency is at maximum 66 per cent and usually runs close to 60 per cent. Inherent tank circuit losses in the amplifier reduce this a bit, so that the plate power output of a representative class-B amplifier may run about 55 per cent. However, in cathode-driven (grounded-grid) service, a portion of the driving power (feed-through power) appears in the plate output circuit and provides a measurable output efficiency of approximately 60 per cent for the stage.

Now, if your maximum input power level is known, as well as plate efficiency, it is easy to determine the power output of the amplifier as well as the power dissipation of the tube (generally known as *plate dissipation*).

If the 2-kW PEP power input condition is chosen and an overall amplifier efficiency of 60 per cent is assumed, the PEP output will be

$$2000 \times 0.60 = 1200 \text{ watts}$$

The remainder of the power ($2000 - 1200 = 800$ *watts*) is consumed in plate dissipation and circuit losses. Tube plate dissipation at 60-per cent efficiency runs close to 800 watts. Circuit losses run from 50 to 100 watts. These figures may add up to a little more than 800 watts of power loss, but a portion of this is accounted for by the plate dissipation attributable to the feedthrough power previously mentioned. So, without doing anything more complicated than a little grade-school math, the general operating parameters of the amplifier have been determined.

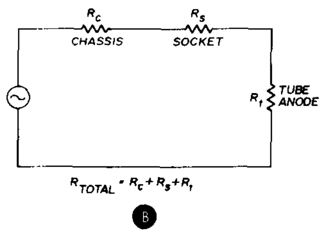
At the 1-kW CW condition, power input is 1000 watts and tube efficiency remains close to 60 per cent, provided certain circuit precautions are taken (these will be discussed later). You can estimate the power output and tube and circuit losses yourself for this power level.

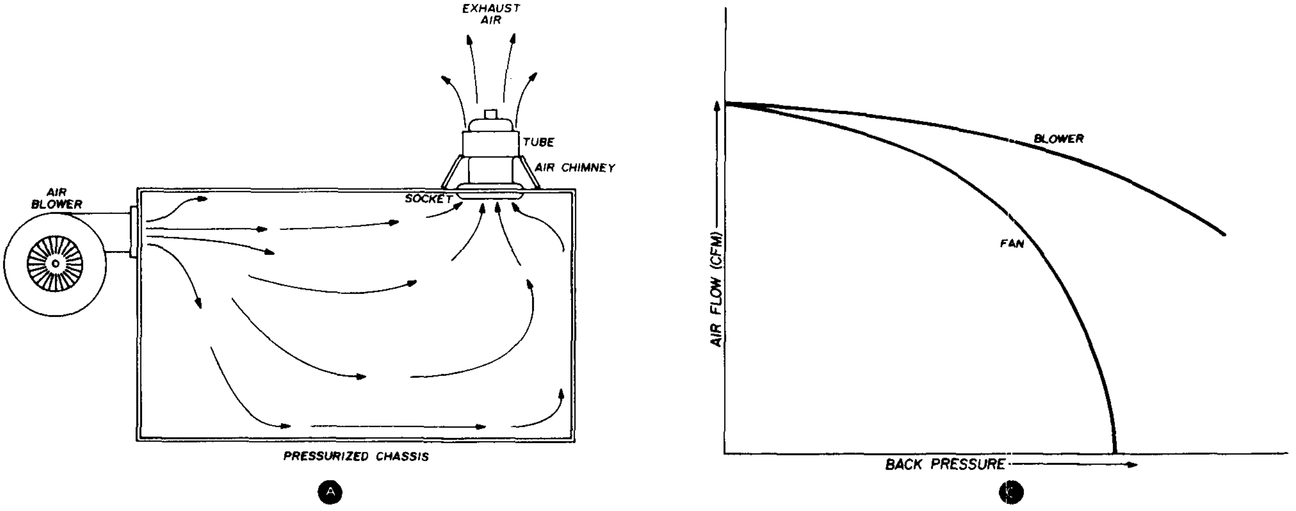
Linear amplifiers and their power supplies have grown sleek and physically smaller in recent years. More efficient components are used and cooling techniques have improved, permitting the amplifier to be squeezed into a compact cabinet with high eye appeal. Some manufacturers and designers, however, have cheated, by skimping on the power transformer or by using an inadequate cooling system that allows the tubes to overheat during extended periods of operation. One old timer, when asked to judge the relative merits of two competitive, widely advertised linear amplifiers, replied, "That's easy! You don't need any test equipment at all. All you do is weigh the amplifiers. Unless one of them has lead fishing weights in it, the heavier amplifier is the toughest and best!"

There is more than a grain of truth in this remark. Attempting to cram a 2-kW PEP amplifier into a shoe box is a time consuming and complicated task, since the problem of getting rid of the heat caused by tube dissipation and circuit losses is a formidable one. Imagine a metal box the size of a 2-kW PEP linear amplifier with an 800-watt bulb burning inside of it! Or consider that a burner on an electric stove may be only 600 watts. This will give you a picture of the amount of heat that has to be removed from a 2-kW PEP linear amplifier during operation to prevent it from burning up.

The human voice, which is the usual modulating device in ham radio, luckily has a low average power level with quite high peak power. Thus, an amplifier designed for voice operation can have a power supply designed for low average power, yet be capable of sustaining full peak power for a short time interval. Many manufacturers count on this low average voice

fig. 2. A representative air cooling system for a ceramic-metal power tube, such as the 8877. A forced-air cooling system is shown in (A); the blower is mounted on the chassis which acts as a plenum chamber. With the chassis airtight, the air is forced past the tube socket, tube base, and out the anode. The chimney is used to direct the air through the finned anode. An electrical analogy of the cooling is shown in (B). The blower is represented by a generator and the various back pressures by the voltage drops across the series-connected resistors. Total back pressure is the sum of the resistances. Representative fan and blower performance is illustrated in (C). The blower efficiency drops as the back pressure is increased, while the fan fails to deliver air at any appreciable back pressure. The ability to overcome back pressure is proportional to the speed of rotation of the blower or fan, plus the physical design of the blades. An inefficient fan allows air to slip around the ends of the blades. It is difficult to determine a good blower or fan, as opposed to a poor one, by intuition. Graphs of blower and fan performance may be obtained from the manufacturers.





power and skimp on the power transformer in an effort to squeeze their amplifier into a small cabinet. But what happens if speech processing is used to raise the average voice level and the amplifier is operated continuously during a DX contest? Or the amplifier is used for RTTY? The power transformer, amplifier, and tubes may not stand up under this added burden. Make sure your design is capable of tough, continuous operation. This is the least expensive approach for the long run.

tube cooling

Once the tube type has been chosen, the next item of business is to adequately cool the tube. The manufacturer's data sheet provides maximum tube temperatures and usually the amount of air required to do the job. What does this entail? An air cooling system is shown in **fig. 2**. It can be compared with a series electrical circuit, wherein the resistance to the flow of air created by the tube and accessories is equivalent to the opposition to current flow provided by resistors. The air resistance (back pressure) is equivalent to the voltage drop across the resistor, and the number of cubic feet of air per minute (cfm) required to overcome the back pressure can be compared with the voltage necessary to force current through the resistors. Back pressure is measured by a manometer and is expressed in terms of equivalent inches of water. Once back pressure and cfm are determined, the blower can be chosen that will force the required air through the system.

Air requirements for some popular transmitting tubes are listed in **table 2**. As an example, the 8877's maximum operating temperature is 250 degrees C. To hold this value, about 22 cfm are required to overcome a back pressure of 0.2 inch of water. This provides an anode dissipation of 1000 watts, more than sufficient for ssb operation at the 2-kW PEP level. The full anode dissipation rating of 1500 watts can be achieved with an air flow of 35 cfm, but at the price of a higher back pressure value of 0.41 inch of water. In passing, it should be noted that axial fans do not like working into high values of back pressure, as **fig. 2C** indicates.

A single 3-500Z tube requires 13 cfm air flow at a back pressure of 0.08 inch of water *per tube*. For two tubes, the *air flow requirement doubles to 26 cfm, but the back pressure remains the same*. Generally speaking, air flow is easy to obtain, but back pressure ability is hard to come by in simple, inexpensive, and relatively noiseless blowers. Amateurs like blowers that don't make noise. Unfortunately, movement of air creates noise, and the higher the back pressure requirement the more air noise that will be created. (This limits the size of a practical, air-cooled, transmitting tube to about 50 kW, above which it would probably require a Volkswagen engine to run the blower and would produce sufficient noise to drive the operator out of the station. Hence, the use of water or vapor cooling in the largest transmitting tubes.)

choice of blower

The most common air impellers are the centrifugal (squirrel cage) blower and the axial fan. The axial fan is the quieter of the two, but does not have the ability to work into a high level of back pressure. The ability of the squirrel cage blower to overcome back pressure is a function of the blower speed in rpm and the diameter of the wheel — the larger the diameter, the lower can be the rpm for a given amount of back pressure. Suggested blower specifications and fan information are included in the data for popular tube types.

It must be remembered, too, that the hypothetical Amateur living at an altitude of 1600 meters (5000 feet), in Denver, Colorado, for example, exists in a world of thinner air than that encountered at sea level and would have to increase the air requirements out-

table 2. Representative cooling requirements for various power tubes.

tube type	cfm	back pressure	blower diameter	rpm
3-500Z	13	0.08	3	1600
(2) 3-500Z	26	0.08[3]	3	3100
3-1000Z	25	0.43	3¾	3000
8874	8.6	0.37	2¾	3100
8875	[2]	0.16	[4]	2800
8877[1]	22.5	0.20	3	3100

Notes 1. For 1000 watts anode dissipation
2. 1600 feet per minute from axial fan
3. In EIMAC SK-410 socket with EIMAC SK-406 chimney
4. Axial fan or blower

The listed values are given in cubic feet per minute and back pressure in inches of water. The impeller information is the centrifugal blower wheel diameter (inches) and motor speed in revolutions per minute. Low-speed blowers are attractive because they create less air noise, but they are unable to work into any appreciable amount of back pressure. As shown, these tubes, regardless of plate dissipation, require a blower speed of about 3000 rpm, except for the single 3-500Z. These data are for operation at sea level and the quantity of air should be increased about twenty per cent for operation at high altitude (Denver, Colorado, for example). The cooling requirements can be verified only by making temperature measurements on the tube seals and the anode. Glass tubes, such as the 3-500Z, can be cooled from the side by an axial fan, but only after tests are made to ensure that the glass envelope temperature remains within specified limits.

lined in the illustration by about twenty per cent to achieve the same degree of cooling.

amplifier enclosure

Once the cooling requirements have been determined, all that remains is to get the cooling air into and out of the amplifier box. Why enclose the amplifier? Aside from cooling requirements, today's electrical specifications require rf harmonic suppression of a high order. This means that the amplifier must

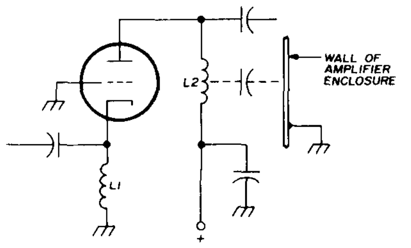


fig. 3. Both input and output chokes of a grounded-grid amplifier can form a parasitic oscillator circuit. Usually the cathode choke, RFC1, has more inductance than the plate choke, RFC2, but stray capacitance between the plate choke and the enclosure can lower the resonant frequency of the plate parasitic circuit until an uncontrolled oscillation can occur. This unwanted oscillation can be cured by removing turns from the plate choke, moving the choke farther away from the metal walls, or by placing a resistor either in series or in parallel with the choke. Means for detecting such parasitics are discussed in the text.

be placed in an rf-tight box and that connections to the amplifier be carefully filtered to prevent unwanted harmonic energy from escaping and blocking out Joe Sixpack's television receiver next door. All amplifiers generate and amplify harmonics of the driving signal; the task is to keep them from harm's way. Proper filtering will do the job.

Ventilation holes can be placed in an rf-tight box, provided they are properly screened. Wires can enter and leave the box provided they are properly filtered. A screened opening should be about twice the size of an unscreened opening to obtain the same air circulation, since the screening material represents nearly 50 per cent coverage of the area. A series of many small holes drilled in the top and bottom of an enclosure will provide ventilation without letting any great amount of rf energy escape, provided the holes are small compared with the harmonic frequency. For high frequency work quarter-inch holes are satisfactory. More smaller holes will work, too. Copper wire screening can be placed over the blower opening if the mating surfaces between screen, blower, and chassis are free of paint so that electrical continuity exists between the various metals.

indicating meters

Several meters are required to properly tune and operate a linear amplifier. At the very least, grid and plate currents should be monitored, and it is convenient to be able to read filament voltage. The grid and plate currents can be read on one meter switched between the appropriate circuits, but the use of separate meters is recommended for ease in tuning. All meters should be checked for accuracy before installation in the amplifier.

Placing the meters in the walls of the amplifier box is bad, since the rf energy can easily escape through the meter case and glass, invalidating the otherwise good shielding of the unit. It is wiser to place the meters on a separate front panel, with the amplifier box supported behind the meter panel. Meter leads are then brought out through appropriate filtering networks.

parasitic suppression

"You don't have to worry about shielding or neutralization. A grounded-grid amplifier just won't oscillate." Right? Wrong. A grounded grid amplifier makes a very good oscillator under certain conditions.

Low-frequency parasitic oscillations. Any amplifier can oscillate in the low-frequency region (200-1500 kHz) by virtue of the interelectrode capacitances of the tube forming some resonant circuit with either the input or plate rf chokes (**fig. 3**). A sure cure for this problem is to change the type of choke, or else place a resistance in series or in parallel with the choke to inhibit oscillation. In the designs discussed here, the inductance of the input choke is very low compared with that of the plate choke, so that oscillation is improbable.

A low-frequency parasite can often be heard in a nearby broadcast receiver as an unsteady carrier or a rough buzz. Or, it can be found when the amplifier is operated with plate voltage (but no excitation) and the controls tuned at random. A small neon lamp is held near the plate lead. If a parasite is present the bulb will ignite with a bright *yellow* glow. The bulb should be held at the end of a dry wooden stick, as dangerously high voltage is present and exposed when the amplifier is operated with the cabinet shielding removed.

Vhf parasitic oscillations. Vhf parasites are created by resonant circuits formed by connecting leads and interelectrode capacitances of the tube (**fig. 4**). They can be suppressed by loading the circuit until oscillation is impossible. A parasitic choke, composed of an inductor and resistor in parallel, will do the job. The suppressor is placed in the plate lead, but a second suppressor is sometimes required in the input circuit.

The suppressor represents a portion of the lead wound up into a coil and shorted by a resistor. At the parasitic frequency there is a large voltage drop across the coil. The resistor acts as an rf load for this

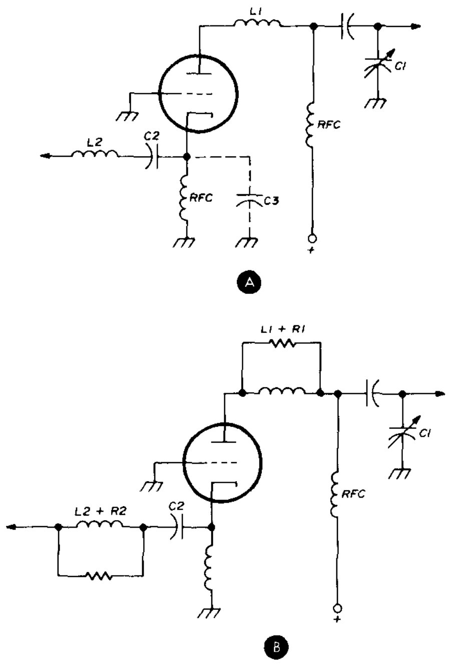


fig. 4. Vhf parasitic circuits in grounded-grid amplifiers are made up of stray input capacitance (C3) plus the inductance of the input and plate leads, L1 and L2, see (A). The parasitic oscillations may be suppressed, as shown in (B), by shunting a portion of the input and/or plate lead with a resistor to load the parasitic circuit. However, this choke must not be too tightly coupled to the plate circuit at the operating frequency or it will dissipate fundamental frequency power and overheat.

voltage drop. If the load is tightly coupled to the tube, oscillation will not take place, but if the load is tightly coupled at the operating frequency, the suppressor will probably overheat and burn up. The number of turns in the inductor must be determined by test so that sufficient inductance exists to do the job, but not enough inductance is used to couple too much fundamental energy into the resistor,

A vhf parasite can be determined by the neon bulb test. The bulb will glow with a bright *purple* color if oscillation is taking place.

High frequency parasitic oscillations. The grounded-grid amplifier can be turned into a splendid oscillator if the input circuit is detuned too far from resonance. The tuning range of the input circuit should therefore be quite restricted. It is a good idea to tune the input circuit to the middle of the Amateur band in use and then forget it. Actual adjustment of this circuit will be discussed in a subsequent article.

High frequency parasitic oscillation can also take place if output power from the amplifier finds its way back into the input circuits. Shielding and filtering (which also reduces vhf harmonic energy radiation) serves to reduce the possibility of high frequency oscillation. In some instances, the grounded-grid amplifier must be neutralized to achieve stability. Luckily, this is generally not required for amplifier operation below 30 MHz but the builder should be aware of the fact that high frequency oscillation at the operating frequency can take place in a grounded-grid amplifier if the proper precautions are not exercised.

summary

So far, we've slogged through a quicksand of physical design problems. Let's sum up what has to be done as far as this aspect of amplifier design is concerned:

1. The amplifier tube has to be chosen for the particular power level desired and must be properly ventilated and cooled.

2. The amplifier has to be placed in an rf-tight box for harmonic suppression and operational stability.

3. Circuit design must ensure that unwanted oscillations do not take place.

The next article in this series will discuss the electrical design of the linear amplifier.

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ham radio

AFC circuit for VFOs

Add a new measure of stability to an existing VFO by incorporating this AFC circuit

The search for frequency stability in Amateur equipment, both commercially manufactured and homebrew, has been one of the longest and most fervently pursued in the history of ham radio. In the earliest days, when King Spark reigned supreme, frequency stability wasn't much of a problem since the damped waves covered a considerable portion of the spectrum. However, as the state of the art advanced (and in those days it was an art), the vacuum-tube oscillator became *re rigeur*, receiver selectivity improved (necessary because of a more crowded spectrum), and the need for greater oscillator stability quickly became apparent. That need has been with us ever since!

In the early 1930s, crystal controlled transmitters became the way to go, with a powerful assist from the Federal Radio Commission (the ancestor of the FCC). You could actually find the same ham at the same spot on the dial (almost) night after night. After all, during the depression, who could afford more than one or two crystals? They cost about two days' pay (about $7.50) each! Crystal control was undeniably a great advance over the self-excited oscillators then in use for transmitter frequency control. By the standards of the day, transmitters became literally "rock stable." Unfortunately, receivers of that period didn't enjoy the same stability. The superheterodyne had become popular, and the instability of its high frequency oscillator (HFO) required a frequent touch of the tuning dial (in much homebrew gear) to keep the desired station audible. Manufactured gear was better, of course, but few Amateurs could afford it. By today's standards these oscillators were pretty crude, but we were dealing with a lot of CW and a little a-m telephony. With the broad bandpass windows in those old receivers, a-m was easily handled, and a changing CW beat note could be tolerated as long as the band wasn't crowded.

Use of the Amateur bands was rapidly increasing, however, and by 1939 or 1940 the trend was away from crystal control toward something known as the ECO, or electron-coupled oscillator. Great strides had been made in stabilizing the same old Hartley and Colpitts oscillators by using higher gain tubes such as the new pentodes, looser coupling between oscillator and load, and a myriad of other tricks that today we take for granted.

Rather than labor the points unduly, it is best to say that today, both VFO and crystal oscillators are in wide use in the ham bands, the more stable crystal oscillator being used for certain receiving functions, MARS frequency assignments, vhf repeater channels, and so forth. The VFO, of course, finds its major application in the HFO of the modern hf transceiver. A new state-of-the-art device, the frequency synthesizer, actually uses both techniques: a reference crystal oscillator of high stability and a particular type

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Front view of the VFO-stabilizing unit. The three LEDs are used to indicate the position of the control voltage within its total range.

of VFO known as the voltage-controlled oscillator (VCO) locked to the crystal oscillator by the phase-locked loop.

Still, the great majority of equipment in use today employs the VFO in one form or another, tube or transistor, Hartley, Colpitts, Vackar. They share one common failing; they *all* drift in frequency to some extent. Even the justly famed Collins PTO can be irritatingly unstable to a frequency-measuring nut.

Individual component variations in an otherwise sound VFO design can, in a production situation, become an annoying source of trouble. A more fundamental problem is that to thoroughly stabilize a VFO takes much time and use of temperature-compensating capacitors. The stabilizing procedure also requires a manufacturer to employ a trained technician for this task. The manufacturer attempts to arrive at a sort of stability "middle ground," or performance that will satisfy the majority of buyers.

what is a VFO

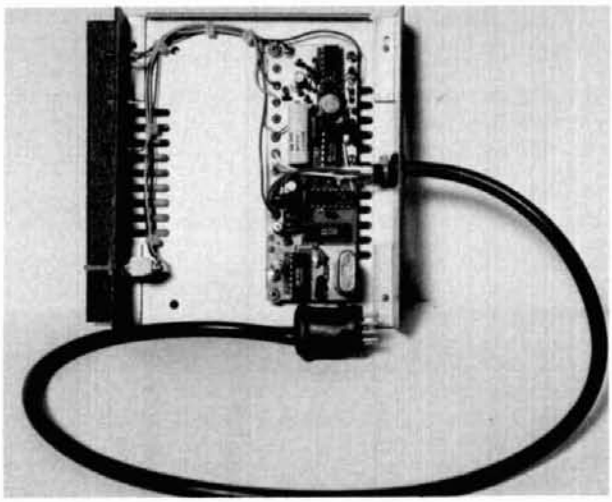
Let's briefly examine the device we're trying to improve so that a little more than the tip of this iceberg becomes visible. Contrary to opinions expressed by some sources, black magic does not play a major role in VFO design. The modern VFO is a marvel of construction. It is usually built in a very rigid box of steel or heavy aluminum, and almost invariably today is a solid-state device even though vacuum tubes may appear elsewhere in the radio. Use of solid-state devices removes from the VFO one of its worst enemies — heat. Most of the VFO circuits in production equipment today can trace their origin to the basic Colpitts circuit, probably because the coil tap in the Hartley introduces added switching complexity and cost.

Even more basic is the real heart of the VFO, the tank circuit or frequency-determining network. *Anything* that causes the oscillatory period of this network to vary (other than deliberate variation) will have an adverse effect on the stability of the VFO. The sole exception to this statement is the AFC circuit. However, this can be considered a form of operator control and, in fact, the operator can alter the VFO frequency through manipulation of the AFC circuitry. Undesirable factors include changes in temperature, change in the load on the VFO, changes in supply voltage, and vibration. An excellent measure of the designer's success is the stability of his VFO. From a manufacturing standpoint, add one more vital point. It must be reproducible! I once built a little VFO that was a marvel of stability. Three other people tried to reproduce it. Their versions oscillated and they were in the correct frequency range — but they drifted. The circuit was not easily reproducible. Therefore, it was useless except to me.

VFO improvement

Many papers have exhaustively covered the trials and tribulations of VFO design, both from the homebrew aspect and from the commercial viewpoint. It seems sufficient to note that most of the hints, techniques, design criteria, and other good and valid information presented in VFO design papers[1,2,3] are basically aimed at reducing frequency drift. We have relatively simple formulas for calculating the component values, but no one has yet come up with a magic formula to make the VFO stable.

The recent article by PAØKSB[4] suggests one solution to the stability problem. **Fig. 1** shows a stabilizing device in block diagram form — a single stage



Inside view of the logic portion of the control unit. The only other components are those installed inside the VFO enclosure. Additional information regarding the changes to the Atlas equipment can be obtained either from the author or by writing to Atlas Radio.

binary counter, a storage element (D-type flip-flop), a clock (low frequency crystal oscillator), and an integrator to drive the control element. **Fig. 2** shows the entire control element, a tuning diode, a bypass capacitor, an isolation resistor, and a small coupling capacitor. Room for these tiny components could easily be found in any VFO I've looked at. As you can see from the size of the coupling capacitor (1 pF), the effect of the VFO (dial) calibration won't be a major one. In all cases so far, minor readjustment of individual band trimmers has easily absorbed the small additional capacitance represented by the control element.

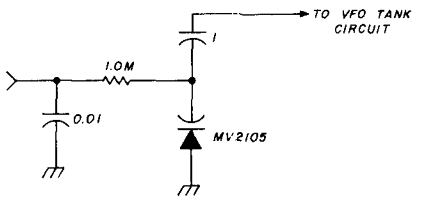


fig. 2. Schematic diagram of the control unit which is installed inside the VFO. Since the added capacitance is very small, any change in frequency is easily compensated for by the VFO's trimmers.

practical AFC package

The actual circuit of the AFC unit is shown in **fig. 3**. There are a few changes from the original. First, in this country at least, there are no 100-megohm ¼-watt resistors commercially available. To my knowledge, the highest value ¼-watt resistor available in quantity in this country is 22 megohms. A series pair of these resistors, along with a 2-μF capacitor (metallized Mylar or polycarbonate) will provide an 88-second integrator time constant, as compared with the 100-second value in the original circuit. The slightly shorter time constant has proven entirely adequate in practice. It must be noted that high-leakage capacitors, such as electrolytics, are unsatisfactory in this application. The added expense of the low-leakage capacitor has to be accepted to obtain a smoothly operating unit.

The circuit likes a low impedance input; this point is mentioned in the original article, but without particular emphasis. As a majority of the radios manufactured in this country have a VFO output impedance in excess of the 50 to 100 ohms which I (and apparently the AFC circuit) consider low impedance, a broadband autotransformer has been added at the input to help the interface problem. Use of this approach has allowed sampling of the VFO output voltage without detriment to either its output amplitude or intrinsic stability. This particular transformer has been examined with a Hewlett-Packard 250B Rx meter and found to be essentially flat from about 2 to over 23 MHz. This is more than adequate for any range of VFO frequencies yet encountered.

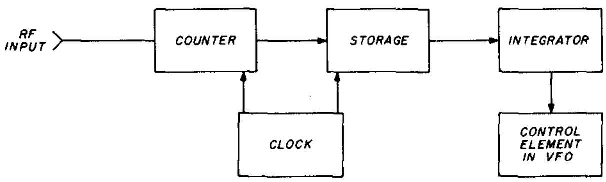


fig. 1. Block diagram of the AFC control unit. The clock controls both the counter and storage sections, determining when to count and when to transfer information to the next stage. The integrator serves to smooth out the information going to the control element, preventing the VFO from jumping back and forth in frequency.

Routine equipment turn-on generally gives a ramp voltage of about one quarter maximum. This, plus normal voltage change due to operation, necessitated some form of metering. An early approach with Atlas equipment involved switching the S-meter. However, this method has since been discarded in favor of three LEDs which, with the addition of a 7406, illuminate upper and lower limit red warning lights when the ramp voltage comes within one volt of either end of its range. In between, a green LED glows. The transition from red to green, at each end of the range, is reasonably abrupt. Only a few millivolts of overlap exist between the red and green LEDs. This metering system provides an adequate Go/No-Go indication which is particularly useful under mobile operating conditions. Trying to read a conventional analog meter under crowded freeway conditions is not exactly conducive to an extended life expectancy! Also, the components of the LED metering system are noticeably less expensive than a meter.

In addition to metering, this approach makes possible construction of the AFC circuitry in a completely separate box without a multitude of connecting wires. The length of the four-wire umbilical isn't critical, allowing the user to position the box for optimum convenience. This concept also reduces the amount of internal work necessary to incorporate the drift-cancelling circuitry. For example, in the case of the Atlas 180/210 series, the device plugs directly into the AUX VFO socket. A 100-ohm resistor at the radio end of the connecting cable prevents any interaction between the AFC system and the VFO in the radio. Internal work on the radio is limited to installation of the control circuitry within the VFO compartment and adding one wire from the control element to pin 1 of the AUX VFO socket (normally unused) to carry the ramp voltage. All other voltages are already present. The digital dial (if used) plugs into this same socket, but, because of the plug construction on

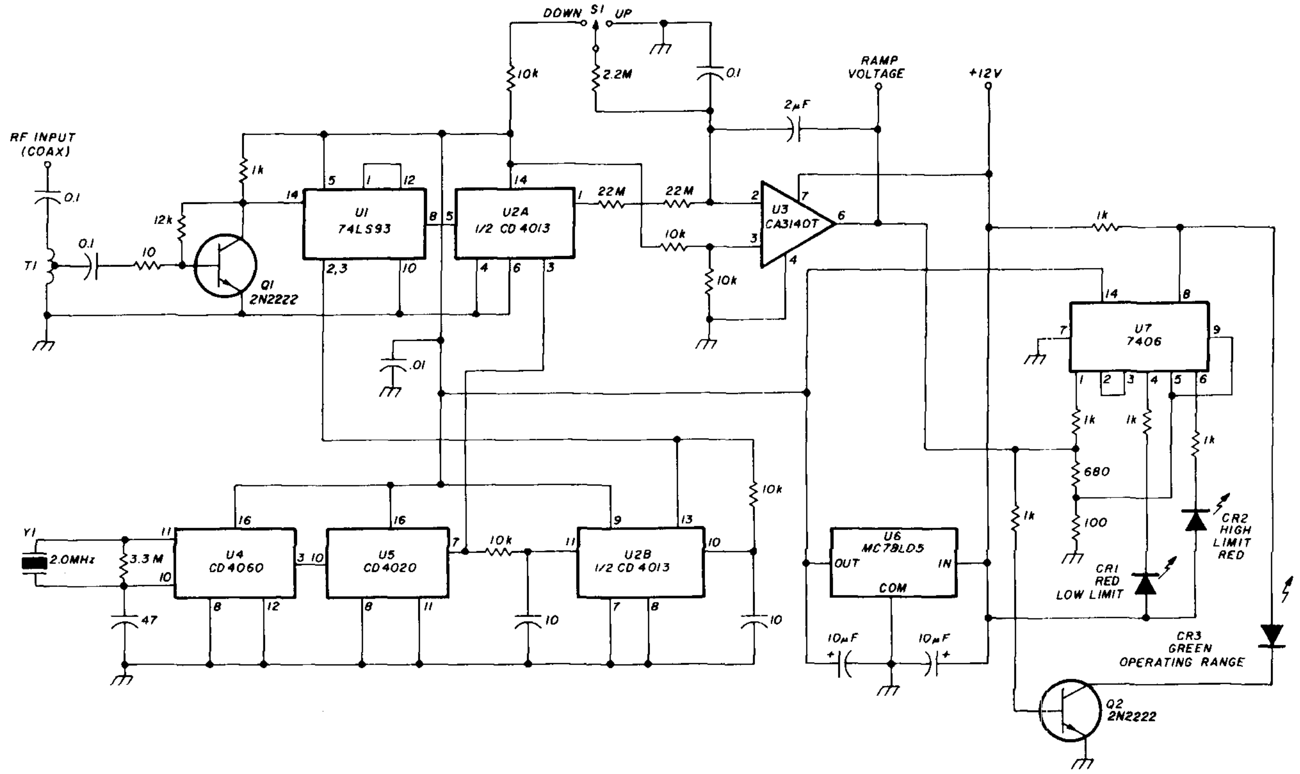


fig. 3. Diagram of the logic portion of the AFC control unit. The transformer is bifilar wound with nine turns of number 26 AWG (0.4-mm) enamel insulated wire over a Q2 ferrite toroid. Core should be 9.4 mm (0.37 inch) outside diameter, 5 mm (0.2 inch) inside diameter, and 4.9 mm (0.19 inch) thick.

both the digital dial and the AFC unit, they piggyback without trouble or interaction. In a transceiver as compact as the 210X, minimal work within the radio is a worthy consideration. Conversely, within the Atlas VFO compartment there is more than adequate room for the control-element components. After having modified several of these radios, total conversion time is less than an hour, including the minor recalibration required.

adaptability

The AFC unit will perform equally well on older tube-type VFOs. It has successfully been installed in Heathkit tube-type VFOs (SB400 series) with no problems. No adverse effects on dial linearity were noted. Since this VFO is not bandswitched, only one recalibration is needed. Even though the piston capacitor is of very low capacitance, more than enough range is available after adding the AFC control element.

Initial examination of an older Swan 350 proved enlightening. Although the schematic shows a solid-state VFO and emitter follower practically identical to the early Atlas units (as well as a tube-type VFO amplifier), maximum rf voltage was only 60 millivolts (at 7 MHz). Other bands weren't much different. This rf level doesn't allow resistive isolation in the pickoff line, because the input voltage to the AFC unit then becomes too low for reliable operation of the counter.

Even with a relatively poor VFO, AFC lockup will occur within a minute. It takes a little time for the integrator to do its thing. A good VFO will normally stabilize within about 15 seconds of turn-on. For example, the Atlas 350-XL with AFC settles down in ten to fifteen seconds, while a relatively poor homebrew unit, the first one on which the AFC approach was tried, took about a minute to stabilize. However, this VFO was so bad that without AFC it would drift noticeably during a two-minute transmission. I threw the whole thing together in a hurry several years ago and never got aroung to taming it. Consequently, it was a natural for AFC experiments. The first time the modified rig was used on the air, I was accused of having bought a new radio. "We don't have to chase you any more!" Constant attention had to be paid to the ramp voltage though, because that VFO never did settle down on its own. About once an hour the ramp voltage would have to be reset because it would be getting perilously close to its range limit. Conversely, the Atlas 350-XL can run all day and nothing has to be reset. The ramp voltage has more than enough range to keep this excellent VFO locked up indefinitely.

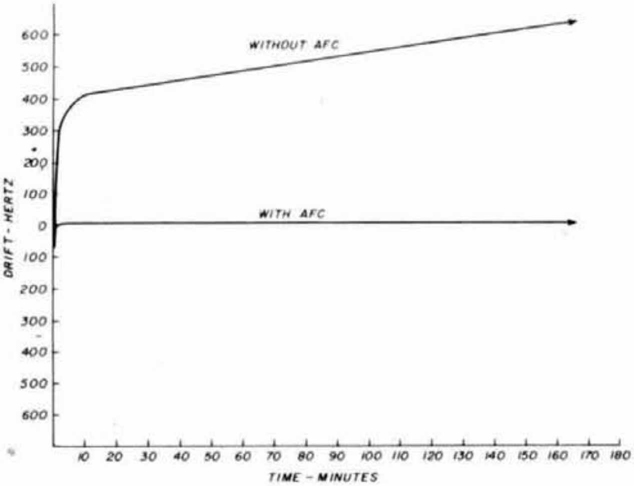


fig. 4. Examples of measured drift from VFOs with and without AFC.

The system of automatic frequency correction described in this article offers an order of magnitude stability improvement. It is not a universal cure for all the ills that may beset a VFO. For example, this circuitry cannot correct the frequency jump that is caused by sticking or misaligned mechanical tuning assemblies or from worn or corroded contacts on a bandswitch or variable capacitor shaft. Sudden frequency excursions of this type look to the circuit like "human intervention," a frequency shift too rapid to be corrected. One demonstration of AFC capability that invariably generates astonishment is connecting a counter capable of reading the VFO frequency to the nearest Hertz to a cold radio which is AFC equipped. When the radio is turned on, the counter will faithfully reveal the rapid initial warmup drift, the sudden stop, and then will remain for hours within about 5 Hz of the original lockup point (see **fig. 4**).

In the near future, buyers of Atlas equipment will be able to order this feature as an option; a retrofit program exists for older Atlas equipment. For those brave souls willing to invade the mysterious innards of their VFO, the system is available as a package; the modification really isn't all that hard to perform. If AFC of the VFO isn't the final answer to outstanding stability in older gear, it must be pretty close to it.

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ham radio

improving antenna accuracy

in satellite tracking systems

A review of the problems encountered in Amateur tracking systems, with suggestions for improving accuracy and operating efficiency

Most users of the OSCAR communications satellites have some form of antenna tracking system for obtaining the greatest advantage from the available transmitter power and for increasing the strength of the received downlink signal. These systems range from simple, manually switched dipoles to elaborate arrays driven in azimuth and elevation under microprocessor control. But most setups fall somewhere in between — often with antennas driven in azimuth and elevation by rotators with separate control boxes manually operated by the same person carrying out the communications — similar to the arrangement shown in **fig. 1**.

This general arrangement is simple, inexpensive, and not particularly difficult to operate when the satellite is low in the sky. On these passes, neither azimuth nor elevation changes very rapidly, and it's a simple matter to update the antenna position from time to time. However, as the satellite gets closer and the elevation angle gets higher, things begin to change much more rapidly, which requires more frequent updating of antenna direction. On some passes activity gets downright frantic as the satellite passes close overhead and the operator finds that the elevation rotator is about to hit the stops and the azimuth angle has suddenly switched from south to north!

Even microprocessor-controlled systems are not entirely immune to these problems. What happens is that the azimuth angle begins to change so fast that the rotator simply can't keep up with it. Add to this the time lag in updating the antenna position in a manually controlled system, and you frequently find yourself in a situation where the elevation control has the antenna pointing straight up (or nearly so) as the satellite azimuth angle suddenly swings 150 degrees or more in a few seconds. So then you're stuck. The azimuth rotator slowly trundles around the 150 degrees (or 210 degrees in the *other* direction if the mechanical stop happens to be in the way), while the satellite merrily recedes into the distance.

Slew rate. The rate at which the azimuth angle changes is called the *slew rate*. What we'd like to do is find a way to anticipate excessive slew rates so we can reduce or eliminate the problem. The first order of business is to find the minimum elevation angle at which the slew rate can exceed the azimuth-rotator rotation speed. This, of course, depends on what kind of rotator you have, but most rotators turn a full circle (360 degrees) in one minute or 6 degrees/second, so we'll use this figure for the sake of argument.

If you ask a dozen different OSCAR users what that critical minimum elevation angle is you'll get as many different answers, but most will probably say it lies between 60 and 70 degrees. Experience in manually operated Az-El control systems does indeed seem to point to a number in this range; but surprisingly, the correct number for OSCAR 7 is an elevation angle of 87.3 degrees — a mere 2.7 degrees from the vertical! The mathematical procedure for arriving at this number is detailed in Apendix A*

*Appendix A is not included with this article but is available on request from the author. Please include a 229 × 305 mm (9 × 12 inch) self-addressed envelope with 28 cents postage. Overseas readers may send one IRC (5 for Air Mail) and omit the envelope.

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along with the procedure for finding the maximum azimuth error (lag) between satellite and rotator as well as the lag duration. For most rotators the critical angle is extremely high. The faster the rotator, the higher the angle.

Obviously, then, the slew rate can exceed the capacity of the typical azimuth rotator for only a very short time on any given pass (in fact, for well under ten seconds at the *maximum*). So why does the slew rate cause so much trouble? We can best answer that by taking a look at just how high the slew rate can go, and in particular by looking at the special case where the satellite passes directly overhead.

Here we have the satellite approaching, let's say, from the southeast. Since it's going to pass overhead, it's coming straight toward us the whole time it's in view, and the azimuth angle doesn't change at all. (Actually, due to earth rotation, the satellite's ground track is slightly curved, and the azimuth angle *will* change a little.) At the instant the satellite passes through the zenith, the slew rate jumps to *infinity*, then becomes zero again as the satellite recedes to the northwest.

In other words, on an overhead pass, the slew rate is infinitely high if only for an instant. If the satellite passes very close by, but not directly overhead, the slew rate will become very high (but not infinite) and will remain high for a longer period of time.

Incidentally, the rate of change of the *elevation angle* is always quite low. For OSCAR 7, the fastest it can possibly change is a little more than a quarter of a degree per second, so this is never a factor we have to worry about.

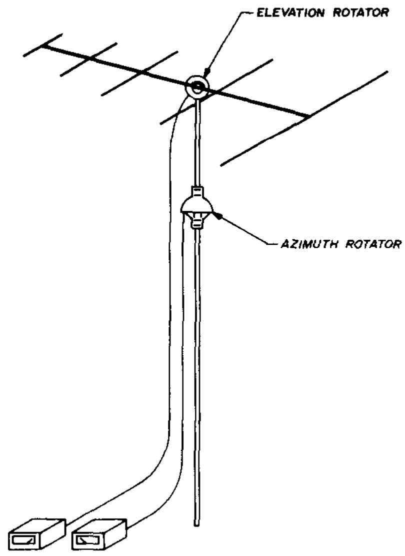

fig. 1. A typical satellite Az-El control system consists of a pair of rotators operated by separate control boxes. Usually the person carrying out the communications must also operate the antenna controls.

As mentioned earlier, OSCAR 7's slew rate can exceed the 6 degree/second speed of a typical rotator for less than ten seconds at the most on any given pass. But even after the slew rate decreases, it still takes a little time for the rotator to catch up. The worst-possible case is again the overhead pass,

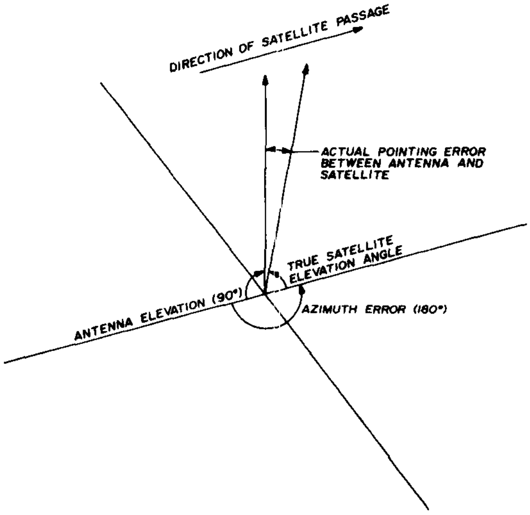

fig. 2. When the satellite is high overhead, or nearly so, the azimuth angle changes quite rapidly and the azimuth error may become quite large. However, with the antenna elevation at 90 degrees, the actual pointing error between antenna and satellite is fairly small.

where the total duration of the lag is 30 seconds. (Appendix A gives details on how to compute the duration of the lag for other passes.)

Pointing error. Even though we can develop a temporary azimuth error of up to 180 degrees, the actual pointing error is not as bad as it might seem. For example, suppose we have an overhead pass and the elevation rotator hits the stops at 90 degrees, as shown in **fig. 2.** The azimuth suddenly swings 180 degrees, and the azimuth rotator direction is now in error by that amount. Even if the elevation control remained at 90 degrees and we waited 30 seconds for the azimuth rotator to come around, the satellite elevation angle would have changed by only about 8 degrees in that time. So the absolute pointing error between antenna and satellite true elevation angle would at most be about 8 degrees, *regardless of the magnitude of the azimuth error*. Once the azimuth rotator comes around to its proper alignment, the elevation rotator can be brought to bear on the satellite in less than 1-1/2 seconds (assuming a 6 degree/second turning rate).

With typical antenna beamwidths of 15 to 30 degrees at the half-power points, an 8-degree pointing error is inconsequential. Judicious operation of the

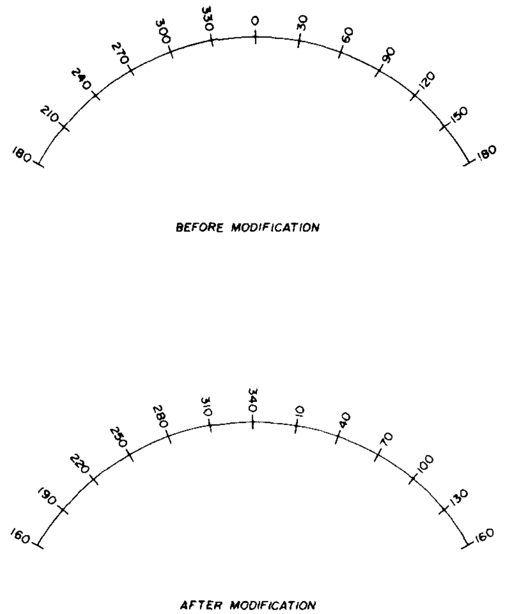

fig. 3. Azimuth rotator dial face as normally supplied for the mechanical stop at 180 degrees and as modified when the stop is placed at 160 degrees.

elevation control can reduce the magnitude of that error even further.

reducing antenna pointing error

Everything we've said so far seems to indicate that the overall problem isn't nearly as great as we thought, so why even bother worrying about it? Well, first of all, the relatively small pointing error described above can be achieved *only* if the antenna is kept on track at all possible times. From a practical standpoint, in a manually operated control system, this simply isn't feasible. If you devote all your time to operating the antenna controls you have no time for communicating.

The end result is that the actual time lags and pointing errors are usually quite large. Nevertheless, a number of ways are available to improve the performance of such a system without putting an increased workload on the operator. Another good reason to attack this problem is that improvement in pointing accuracy allows the use of higher-gain antennas, which have inherently narrower beamwidths. This is especially beneficial when the antenna is under some form of automatic control.

Planning. A good rule in manually operated systems is to plan each pass ahead of time. Check to see how high the elevation angle will be to locate any potential trouble spots. During the pass, instead of playing a constant game of catch-up with the antenna, lead the satellite by half the interval between antenna corrections. For example, if you normally reposition the antenna every two minutes, then each time you move the antenna, set it to where the satellite will be one minute from that time. In this way the antenna will be a little ahead half the time and a little behind half the time, and the average pointing error will be reduced to *half* of what it would have been otherwise.

Mechanical considerations. Consider the azimuth at which the mechanical stop on the azimuth rotator should be placed. Since most rotators turn only 360 degrees, the ideal location for the stop would be at the azimuth corresponding to the point where the satellite first comes into view on a pass that will take it directly over the ground station. (For most locations, this would be an azimuth of about 162 degrees for ascending passes, or about 15 degrees for descending passes.) In this way there would never be any interference from the stop for the type of pass (ascending or descending) you selected for its location, and interference to the other type of pass would be no worse than if you did nothing at all.

It may seem that some rotators require the stop to be oriented to a particular direction (usually south). But this isn't true; you can orient the antenna to any direction you like on the mast. All that's required is that, once you've reset the antenna, you draw a new dial face for the position indicator. **Fig. 3** illustrates an example where the regular dial face of an indicator requires that the stop be at 180 degrees and as modified for the stop at 160 degrees.

You might also try to obtain or modify a rotator that will turn more than 360 degrees. One that will turn 390 degrees with stops set at 180 degrees and 210 degrees, would work very nicely for both ascending and descending passes. As shown in **fig. 4**, if rotation starts from the 180-degree stop, the antenna can move clockwise a full turn through 180 degrees,

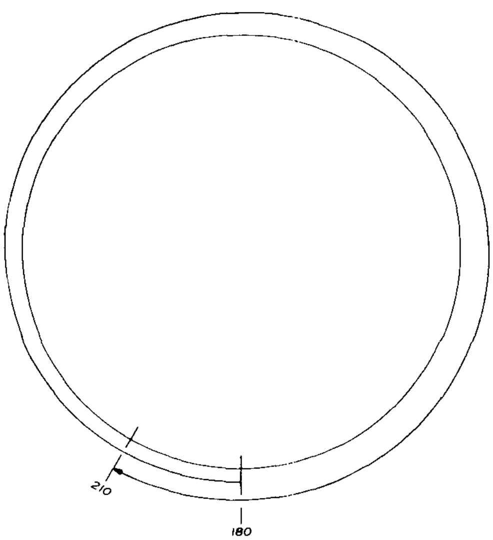

fig. 4. Turning pattern of rotator with 390 degree rotation

finally hitting the other stop at 210 degrees. These stops wouldn't have to be mechanical stops in the rotator but could be electrical stops in the control box or appropriate software if the system is under microprocessor control. Some sort of indicator (a pilot light, perhaps) could be used to let you know when the rotator is inside the extra 30 degrees, so you'll know which way to turn it if you're getting ready to track the satellite's initial azimuth as it comes over the horizon.

antenna elevation inversion system

Although most elevation control systems are set up with stops at 0 and 90 degrees, there's no law that says they have to be that way. And in most cases there is no physical reason that limits the equipment to this range. Remember the overhead pass? Suppose you don't touch the azimuth control, but when the antenna elevation reaches 90 degrees you let it keep right on going over on its back. All of a sudden you've eliminated a whole slew of problems. Instead of frantically chasing after the azimuth control, just ignore it, and every once in a while give the elevation control a slight touch up. Even when the pass isn't directly over head,much time and effort can be saved on high passes by inverting the antenna at the proper moment.

Antenna motion resulting from an elevation inversion system of this type is illustrated in **fig. 5**. Manufacturers of optical and electronic tracking equipment often incorporate this feature into their

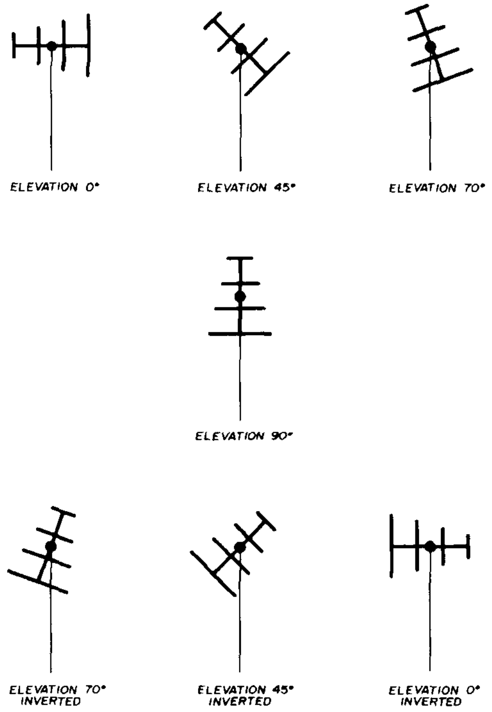

fig. 5. Antenna movement with an elevation inversion system. No azimuth change is made in these illustrations.

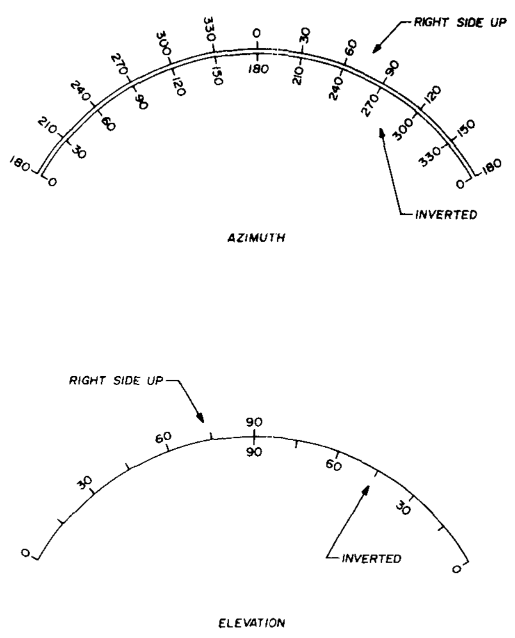

fig. 6. Azimuth and elevation dial faces for an antenna elevation inversion system. Numbers on the normal side of the dials are black; those for antenna inversion are red. When elevation dial is in the red, the corresponding azimuth red scale is used, and *vice versa*.

products, although they sometimes refer to inversion as "dumping."

One thing you must do when using this type of system is to modify the azimuth and elevation control dial faces. Otherwise, when the antenna inverts, the azimuth dial will be 180 degrees off. It's normally a simple matter to extend the scale on the elevation control, and an example of how to label both faces is shown in **fig. 6**. The numbers on the normal (right-side up) side of the elevation dial are black and are red on the side that shows the antenna inverted. The azimuth control has its normal set of black numbers with an inner set of red numbers that are 180 degrees opposite in value to the black ones. When the elevation control pointer is in the black, you read from the black scale on the azimuth dial also. When the elevation dial is in the red, you read from the red scale on the azimuth control. It probably wouldn't hurt to have an indicator light illuminate when the antenna is inverted.

Renumbering dial scales isn't as hard as you might think. Dry transfer lettering works very nicely, and, if you want to be able to return the dials to their original condition, the new scales can be put on paper templates and attached with rubber cement. They can be peeled off later without damaging the original dial face.

systems under microprocessor control

Many of the ideas discussed so far, although especially suited for systems under manual control, are applicable to automatic control systems. Let's take a look at some factors that again affect both types of systems but are of special interest in systems under microprocessor control.

We already know that slew rates vary widely, from a small fraction of a degree per second up to a hundred degrees per second and higher. Most rotators, though, have only one speed, so you have to constantly turn it off and on to keep it on track. These start/stop operations put a lot of wear on the rotator, and the very nature of the procedure prevents the antenna from being exactly on the satellite at all times. Ideally, we'd like a rotator with a speed that varies according to the satellite slew rate, since this would produce a smooth and continuous motion with no pointing error.

Variable-speed rotators. Finding a variable-speed rotator suitable for your station would probably be no easy task, although you could build one from scratch with an appropriate motor. On the other hand, if your rotator is driven by a synchronous ac motor, it would be a simple matter to build a variable-frequency ac power supply that would in turn vary the motor speed. The microprocessor would select the correct frequency to produce the desired rotator speed. This, of course, would require calculating the slew rate in addition to the azimuth and elevation angles. The procedure for doing this is given in Appendix A.

Even the operator of a manually controlled system might find a variable-speed control handy, as it could reduce his workload to a certain extent. Commercially made speed controls, operated by a small joy stick and capable of handling motors of up to 15 watts, are available through a number of amateur telescope dealers. They are advertised as "drive correctors." You can save yourself some money, though, if you're willing to build you own.

It's even easier to vary the speed of rotators driven by dc motors, but whether the rotator is ac or dc, you'll have to invest some time in designing the interface circuitry between the control system and the microprocessor.

Microprocessor interface circuitry. Regardless of whether the rotator speed is variable or fixed, there will be times when the slew rate exceeds the rotator's capability. To reduce the complexity of the microprocessor software, the computer would calculate the slew rates for the pass ahead of time and store the information concerning the time period (if any) during which excessive slew rates will occur. Then, during actual tracking, the microprocessor would instruct the antenna to begin leading the satellite by a few degrees just before it gets to the bad area. The end result would be a worst-case pointing error of no more than 4 or 5 degrees, and then only for five or ten seconds.

There are, of course, many things the software should take into consideration; unfortunately we can't discuss all of them here. But whatever you do, make sure that the software knows where the rotator's mechanical stop is, and make sure it knows how to minimize interference from the stop.

operational procedure for a microprocessor-controlled tracking system

Now that we've discussed a number of the factors in developing a system, let's see how an interactive microprocessor-controlled tracking system might operate. The software would be designed for fully automated operation while at the same time incorporating interrupt capability to permit the operator to take control when desired.

The tracking program could be called up by typing in the word **TRACK**. Once activated, the program would accept additional commands. Entering **NEXT PASS** for example, would cause the computer to check the station's digital clock, compute the starting time of the next pass for each of the active satellites, and flash a message similar to the following on its display screen:

28 OCT 78 16:35 UTC

SATELLITE	MODE	NEXT PASS	TIME TILL START
Oscar 7	A	17:04	:29
Oscar 8	J	16:21	IP
Oscar 9	B	16:44	:09
RS2	A	17:51	1:16
RS4	A	16:50	:15

LENGTH	HIGHEST ELEVATION	DIR	PRIMARY COVERAGE
9	15	A	WEST
12	32	D	EAST
21	88	A	OMNI
4	6	D	NE
18	76	D	OMNI

which satellite?

Suppose, as in example above, you decide to work OSCAR 9, which has a Mode B ascending pass starting in nine minutes and lasting for 21 minutes. (Since it will be an overhead pass, the computer shows the coverage to be omnidirectional.) You would type in **OSCAR 9**, to which the computer would reply:

PREPARE TO TRACK?

Upon receiving a response of **YES**, the microprocessor would make calculations to determine if the azimuth slew rate will become excessive and would store the information regarding that part of the pass. It would then calculate the initial azimuth angle and swing the antenna to that point on the horizon (0 degrees elevation). As this operation is being carried out, the microprocessor would check the clock and display on the screen:

READY OSCAR 9 MODE B
T MINUS 8:43 AZ 163 EL 0

Following this, the computer would update the T time (**TIME TILL START**) every second as it counted down from 8 minutes, 43 seconds. You'd then be free until T minus 15 seconds, at which time the computer would begin emitting a persistent "beep-beep-beep" to alert you of the approaching satellite acquisition. At T minus zero the beeping would stop and the screen would flash:

TRACKING OSCAR 9 MODE B 21:00 REMAINING
AT T PLUS :00 AZ 163 EL 0

As the pass progresses the computer would count down the time remaining while counting up the T time and would continually update the azimuth and elevation numbers so that you'd know where the antenna is pointing at any given moment. A few minutes before the satellite reaches the point where the slew rate becomes excessive, the microprocessor would instruct the antenna to begin leading the satellite by a few degrees. The pointing error will therefore be kept to a minimum, and a few seconds later the antenna will be right back on target. When the pass ends, the screen flashes:

PASS COMPLETE AT T PLUS 21:00 AZ 354 EL 0

At this point you can shut down the station if you're finished. Or, if you wish to try another pass, you can enter **NEXT PASS** to obtain an updated list of upcoming passes. Perhaps you're interested in what's available the next day. In that case you'd enter **ALL PASSES 29 OCT** to obtain a complete list. Or maybe you're interested only in Mode J for the next afternoon, so you'd type in **MODE J PASSES 29 OCT PM** to obtain a list of only the passes in which you're interested.

An interrupt capability would be incorporated into the software so you can stop in the middle of a pass if desired. For example, during the OSCAR 9 pass described above, suppose you communicate for ten minutes, then decide to try a different satellite. You'd type in **STOP**, which halts the tracking of OSCAR 9, followed by **NEXT PASS RS4**, to which the computer replies:

28 OCT 78 16:54 UTC

SATELLITE	MODE	NEXT PASS	TIME TILL START
RS4	A	16:50	IP

LENGTH	HIGHEST ELEVATION	DIR	PRIMARY COVERAGE
14	76	D	OMNI

PREPARE TO TRACK?

This tells you that the pass for the Soviet satellite RS4 is in progress (IP), having begun at 16:50, with 14 minutes remaining in the pass. Furthermore, it's a descending Mode A pass with a fairly high elevation angle and more or less omnidirectional coverage. In reply to the computer's question, you might type in **YES**.

The computer would immediately bring the antenna to bear on the satellite at its current position in the sky and begin to track RS4 as it flashed:

TRACKING RS4 MODE A 14:00 REMAINING AT
T PLUS 04:00 AZ 328 EL 11

The possibilities are unlimited, but the examples we have looked at should give you a good idea of the convenience that appropriately designed software can bring to a station that has a microprocessor-controlled satellite tracking system. You have a free hand in programming all the features that suit your situation.

summary

We began by discussing problems encountered in satellite tracking and described how some of them can cause serious pointing errors in both manually and automatically controlled systems if corrective measures are not taken. A number of approaches were covered for improving accuracy. A major concern is that, if at all possible, we reduce the workload on the operator (who would prefer to spend his time communicating rather than operating antenna controls), and we have presented some ways to do just that. Finally, we looked at some special problems associated with microprocessor-controlled systems and discussed some software features that should be included in the finished system.

We concluded by looking at an example of an operational procedure for a microprocessor-controlled tracking system. Here we discussed how the computer could provide detailed data on various satellites as well as controlling the tracking antenna in a convenient and accurate manner.

A number of hams have incorporated many of these features into their stations. The sky's the limit when you begin to work on your own system.

ham radio


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20M — 13.8-16.0 MHz
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diode noise source

for receiver noise measurements

Construction of a temperature-limited diode noise source for accurate automatic and manual noise-figure measurements to 500 MHz

The saturated temperature-limited thermionic diode has been used extensively for measurement of receiver noise figures in the high-frequency, vhf, and low uhf regions. Its characteristics are predictable and repeatable, and it may be used either in conjunction with an automatic noise-figure meter or with its own power supply and indicating meter for manual noise-figure measurements.

A recent article[1] described the use of diode noise sources with automatic noise-figure meters and indicated that many homebuilt sources could be used

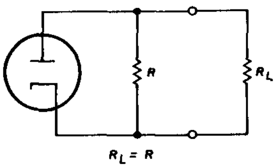

fig. 1. Basic circuit of a temperature-limited diode noise source.

with the Hewlett-Packard models 340A, 340B, and 342A. The article also stated that most, if not all, of these noise sources were unsatisfactory because of their high VSWR at frequencies above 250 MHz or so. Described here is a noise source, using the Sylvania 5722 diode, which is patterned after the Hewlett-Packard Model 343A VHF Noise Source, and which appears to be comparable to that commercial unit at frequencies up to at least 450 MHz.

The noise source may also be used to make manual noise-figure measurements by either the twice-power or Y-factor method. Either technique requires a fixed plate supply and a variable filament supply, with an appropriate plate-current meter. In the past it has been normal practice to vary the diode filament voltage by means of a small variable auto-transformer or a power rheostat. Such control devices change voltage in discrete, albeit small, steps. This, coupled with voltage changes in the primary power source, make it difficult to establish and then to maintain the desired diode plate current. W6GXN described an improved power supply which minimized the problem.[2] This article will present an updated version of his approach, using modern solid-state techniques.

diode noise source

To understand how a temperature-limited thermionic diode is used as a noise generator, we must start with the basic concepts of noise power. A resistance at a temperature other than absolute zero generates across its open-circuit terminals a voltage which is caused by the random motion of free electrons. This noise voltage, e_n, is infinitely broadbanded and defined by the equation

$$e_n^2/B = 4kTR \text{ volts}^2/\text{unit frequency bandwidth} \quad (1)$$

where

k = Boltzmann's constant 1.374×10^{-23} joule/°K
T = absolute temperature in °K
R = resistance in ohms
B = bandwidth, in Hz, of device under test

Since our treatment of noise deals with receivers or amplifiers of finite bandwidth, **eq. 1** is usually written as

$$e_n^2 = 4kTRB \quad (2)$$

When resistance R is connected to a matched load resistance, R_L (equal to R), maximum transfer of noise power will result. The noise power, P_n, dissipated in the load will be

$$P_n = \frac{e_n^2}{4R} = \frac{4kTRB}{4R} = kTB \quad (3)$$

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Eq. 3 defines the *available noise power* from resistance R.

Although its derivation is beyond the scope of this article, effective input noise temperature, T_e, is defined as

$$T_e = \frac{T_{ih} - YT_{ic}}{Y-1} \qquad \textbf{(4)}$$

where

T_{ih} = hot input noise temperature in $^{\circ}K$
T_{ic} = cold input noise temperature in $^{\circ}K$
Y = ratio of hot output noise power to cold output noise power

In the case of a temperature-limited thermionic diode (**fig. 1**), shot-noise current, i_n, can be determined from the equation

$$i_n^2 = 2qIB \qquad \textbf{(5)}$$

where

q = electronic charge, 1.600×10^{-19} *coulomb*
I = dc plate current in amperes
B = bandwidth, in Hz, of device under test

The total available noise power, P_n, from a diode noise source is the sum of the diode shot-noise power and the terminating resistor noise power. From **eqs. 3** and **5**,

$$P_n = \frac{R}{4}(2qIB) + kTB \qquad \textbf{(6)}$$

This equation can be factored and rearranged as

$$P_n = kB\left[\left(\frac{q}{k}\right)\left(\frac{IR}{2}\right) + T\right] \qquad \textbf{(7)}$$

Since

$$\frac{q}{k} = \frac{1.600 \times 10^{-19}}{1.374 \times 10^{-23}} = 11644.8 \qquad \textbf{(8)}$$

then

$$P_n = kB(5822IR + T) \qquad \textbf{(9)}$$

If, in **eq. 3**, the temperature is T_{ih}, then

$$kT_{ih}B = kB(5822IR + T) \qquad \textbf{(10)}$$

or

$$T_{ih} = 5822IR + T \qquad \textbf{(11)}$$

where T_{ih} is the noise temperature of the noise diode with its load resistance, R, at a temperature of T. Thus when I is zero, T_{ih} is equal to T.

The excess noise ratio, ENR, is defined as the ratio of the available noise power at temperature T in excess of that available at a standard temperature (T_o) to the available noise power at T_o, and is expressed as

$$ENR = \frac{kTB - kT_oB}{kT_oB} = \frac{T - T_o}{T_o} \qquad \textbf{(12)}$$

At a standard reference temperature of $290^{\circ}K$,

$$ENR = \frac{T-290}{290} = \frac{T}{290} - 1 \qquad \textbf{(13)}$$

or

$$ENR = \frac{T_{ih}}{290} - 1 = \frac{5822IR + T}{290} - 1 \qquad \textbf{(14)}$$

$$= 20.08IR + \frac{T}{290} - 1$$

Since the term $\frac{T}{290}$ is the noise power contribution of the load resistor, if a temperature of $290^{\circ}K$ is used for the load, **eq. 14** reduces to

$$ENR = 20.08IR \qquad \textbf{(15)}$$

If R equals 50 ohms and I is expressed in *milliamperes*,

$$ENR \doteq I \qquad \textbf{(16)}$$

and

$$ENR_{dB} \doteq 10 \log I \qquad \textbf{(17)}$$

Note that this commonly used equation is predicated on the temperature of the load resistance being 290°K (17°C or 62.6°F). In actual practice, the load resistance temperature may be as high as 310°K (37°C or 98.6°F). In this case, the excess noise ratio will be lower, yielding receiver noise-figure measurements which are higher than the true noise figure. The corrections for these errors are plotted in **fig. 2**.

A schematic diagram of the actual temperature-limited diode noise source is shown in **fig. 3**. You will note that the output circuit is considerably more complicated than those used in most of the previously published designs.[2-5] It is this output circuit, properly arranged physically, which makes this noise

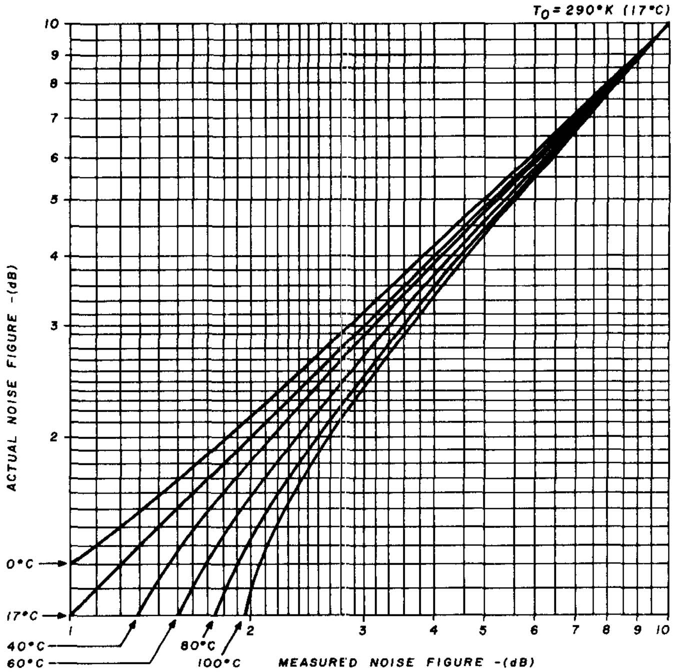

fig. 2. Temperature corrections for a temperature-limited diode noise source (courtesy Hewlett-Packard Company).

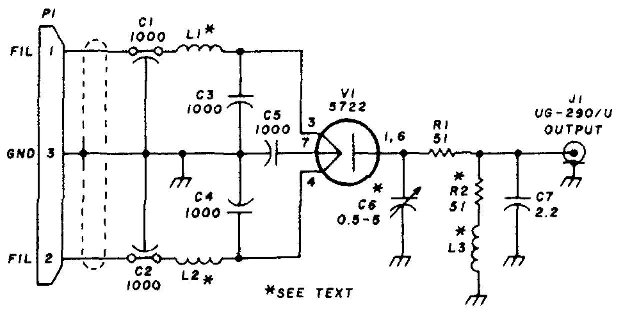

fig. 3. Schematic diagram of the diode noise source. Pin connections shown at P1 are those for a Cannon WK-5-22C-5/16 connector used to mate with a Hewlett-Packard 340B, 342A, or modified 340A Automatic Noise Figure Meter.

source comparable to the Hewlett-Packard model 343A.

The diode output circuit must theoretically satisfy two requirements: it must present a 50-ohm impedance (1:1 VSWR) to the receiver connected to the output connector, and it must present a load to the diode which results in a constant *ENR* over the usable frequency range. In practice, however, it appears that these two conditions cannot be satisfied concurrently, and, as in all designs, a compromise must be reached. The compromise in this case is to keep the VSWR as low as possible and to permit the *ENR* to change over a portion of the frequency range. The rationale for this is simply one of a known factor versus an unknown. If the VSWR is other than 1:1, the mismatch loss between the noise source and the receiver will be indeterminate. (Although noise power obeys all power-transfer laws, noise is random in phase; therefore the loss is ambiguous rather than known.) On the other hand, it is possible to determine the *ENR*, although not to any precise degree of accuracy, so that measurements using a known value are possible.

An expanded schematic of the diode output circuit appears in **fig. 4**. The numbered components correspond to those shown in **fig. 3**, while those having letter subscripts represent the distributed reactive components, as follows. C_{pf} is the plate-to-filament capacitance of the tube; L_p is the series inductance of the tube structure and tube pins; L_s is the series inductance present in R1 and the plate pin connectors; and C_s is the shunt capacitance of the tube socket and/or C6.

If R1 and L3 are both replaced by shorts, and C7 is removed from the circuit, there will be a steep rise in the *ENR* to a resonant point, as shown on curve **A** of **fig. 5**. The VSWR at 432 MHz with this configuration will be about 3.5:1. K2PEY, in his circuit,[3] damped out this resonance by adding a 51-ohm series resistor (R1), resulting in curve **B**. However, the VSWR was still approximately 1.5:1 at 420 MHz according to his article, and about 1.8:1 at 432 MHz by measurement.

W8BBB modified the circuit further by adding L3 in series with R2. Since his article provided no figures on *ENR* or VSWR,[4] there is no basis for its comparison with the earlier configurations or with the one presented at this time.

Incorporating R1, L3, and C7 in the circuit results in curve **C** of **fig. 5**. It can be seen that the *ENR* has a rising characteristic with frequency, but the VSWR is less than 1.2:1 at 432 MHz. Thus, we have achieved a low VSWR, but at the cost of having an *ENR* which changes with frequency. As previously stated, however, we know the *ENR* to a fair degree of accuracy and we have minimized the indeterminate mismatch loss.

The noise source is enclosed in an 83 x 54 x 41 mm (3-1/4 x 2-1/8 x 1-5/8 inch) aluminum utility box. Two methods of mounting and making connections to the 5722 noise diode are presented, both of which have proved to be equally usable.

If suitable VSWR-measuring equipment is available (either a network analyzer or a slotted line and SWR indicator), the method shown in **fig. 6** is preferable, since it permits optimization of the noise-source VSWR by adjustment of capacitor C6. The tube is mounted by means of a 19-mm (3/4-inch) diameter wrap-around plastic clamp positioned so that the cable tube pins pass through a 16-mm (5/8-inch) diameter hole in the shield plate. The cable clamp is attached to the enclosure by means of a metal standoff post. Connections to the tube are made via contacts which have been removed from a 7- or 9-pin miniature tube socket and slipped over the appropriate tube pins.

If you are unable to measure the VSWR of the noise source, use a standard 7-pin, saddle-mount, mica-filled phenolic tube socket, from which the center ground post and the contacts for pins 2 and 5 have been removed; mount the socket on the shield plate. Do not use a black Bakelite or ceramic socket, or one which has a shield base. The added capacitance of the mica-filled phenolic socket is just about optimum at 432 MHz, so capacitor C6 can be omitted.

Other than the mounting and the elimination of C6, the socket method of construction follows that

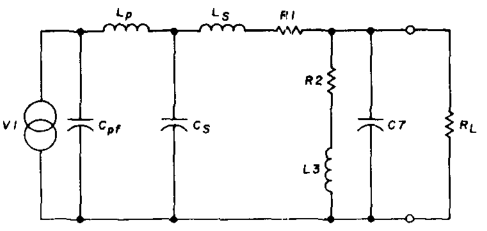

fig. 4. Expanded schematic diagram of the diode-source output circuit. Numbered components correspond to those in fig. 3. Components with letter subscripts are explained in the text.

shown in **fig. 6**. In either case, mount the tube so that pins 3 and 4 are in a plane perpendicular to the bottom of the housing. To reduce the plate lead inductance, the contacts for pins 1 and 6 should be bent at right angles toward one another and soldered together. The usual vhf wiring techniques are required. Disc capacitors C3, C4, and C5 must be soldered directly between the tube pin contacts and the mounting plate, with lead lengths at an absolute minimum. Rf chokes L1 and L2 are each ten turns of no. 26 (0.4-mm) or no. 28 (0.3-mm) enamel-covered wire, close spaced and air wound to approximately 0.1 inch (2.5 mm) diameter.

Although R2 is shown as a nominal 51-ohm resistor, its optimum value is 50 ohms. It is suggested that a resistor as close as possible to 50 ohms (measured on a resistance bridge or accurate digital multimeter) be selected. Five per cent, 1/4-watt *carbon-film* resistors are recommended for both R1 and R2; the application of heat when soldering to the necessary short leads of ordinary composition resistors generally changes their values.

One lead of R2 is wound to form L3, a three-turn coil 2.5 mm (0.1 inch) in diameter and 5 mm (0.2 inch) long. Except for this lead, all other connections to the plate of V1 or to J1 must be made virtually leadless. C7, R2/L3, and C6 (if used) are all grounded to a solder lug attached to one of the mounting screws for J1.

A shielded, two-wire cable, brought out of the housing through a neoprene grommet, is used to connect the noise source to either a Hewlett-Packard automatic noise-figure meter or to the noise-source power supply unit. If the noise source is to be used with a noise-figure meter, the appropriate pin connections to a Cannon WK-5-22C-5/16 connector are shown in **fig. 3**. Otherwise, any convenient pair of mating connectors can be used on the noise-source cable and the power supply. Obviously, the male of the pair should be attached to the cable so that high voltage is not exposed on the power-supply connector.

The 5722 diode generates a considerable amount of heat, which can affect the resistance values of R1 and R2. To minimize the heat within the housing, discard the U-shaped part of the utility box that mates with the half shown in **fig. 6**. In its place, use a similar piece made from perforated aluminum stock. This will retain the integrity of the shielding and will allow heat convection from within the enclosure.

If VSWR-measuring equipment is available and C6 has been included in the circuit, the VSWR of the noise source can be optimized. This can be done without applying filament or plate power to the diode. **Table 1** shows the VSWR of several configurations of the noise source, with and without a tube socket. The frequency at which C6 should be adjusted depends on your preference. Adjusting C6 for minimum VSWR at 432 MHz results in a VSWR of less than 1.2:1 over the entire usable range of the noise source. If the VSWR adjustment is made at 222 MHz, the VSWR increases to slightly less than 1.3:1 at 432

CURVE	R1	L3	C7	VSWR AT 432 MHz
A	0	0	∞	3.5:1
B	51	0	∞	1.8:1
C	51	≈10nH	2.2pF	< 1.2:1

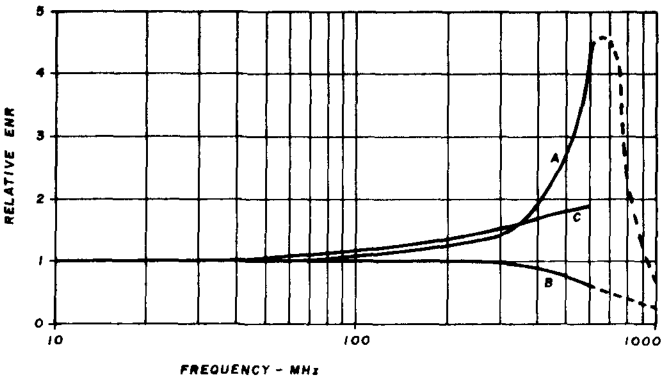

fig. 5. Relative *ENR* of a diode noise source for various configurations of the output circuit shown in fig. 4. Ideally, the *ENR* should be unity over the entire frequency range. Curves A and B have been calculated, thus the dashed portions above the maximum usable frequency of 600 MHz. Curve C is representative of the Hewlett-Packard model 343A and of the homebuilt noise source up to at least 450 MHz.

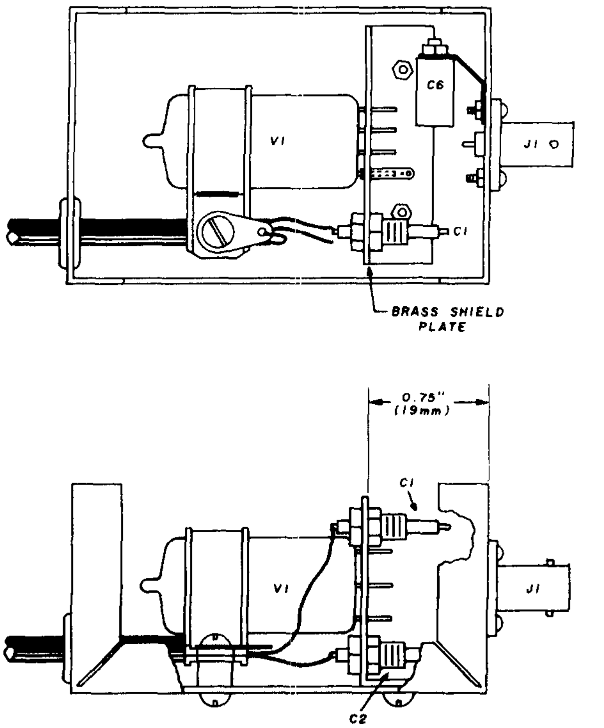

fig. 6. Physical layout of the noise source. The shield plate on which C1 and C2 are mounted is made of brass with a 16-mm (5/8-inch) diameter hole to clear the tube pins or to accommodate a 7-pin miniature tube socket (see text).

MHz and is only 1.1:1 at 145 MHz. Since noise-figure measurements at 432 MHz would appear to be of greater import than at other frequencies within the range of the noise source, it is recommended that the VSWR be optimized at that frequency. When so adjusted, the VSWR at the other amateur frequencies appears to be at least equal to that of the Hewlett-Packard model 343A.

For those amateurs interested in tweaking to the diode noise source for manual noise-figure measurements appears in **fig. 7**. It consists of an adjustable, regulated switching supply for the filament of the 5722 noise diode, supplied by transformer T1, and a simple half-wave rectified dc plate supply. High voltage is obtained from T2, a 12.6-volt transformer which has its low-voltage winding connected to the secondary of T1, thereby providing approximately 115 volts ac for the plate supply.

table 1. VSWR measurements for several configurations of diode noise sources. See text and fig. 3 for description of configurations (second column).

unit	configuration	30 MHz	50 MHz	145 MHz	VSWR 222 MHz	432 MHz	500 MHz	550 MHz
1	no socket, C6 and L3 optimized at 432 MHz	1.04	1.07	1.12	1.15	1.03	1.06	1.14
2	no socket, C6 and L3 optimized at 432 MHz	1.05	1.09	1.20	1.19	1.03	1.12	1.23
2	no socket, C6 optimized at 300 MHz	1.05	1.06	1.08	1.08	1.16	1.27	1.37
2	no socket, C6 optimized at 222 MHz	1.04	1.04	1.04	1.04	1.29	1.41	1.51
3	with socket, C6 omitted, C7 optimized at 432 MHz	1.04	1.06	1.09	1.14	1.05	1.15	1.25
4	with socket, C6 omitted, C7 optimized at 432 MHz	1.06	1.10	1.18	1.22	1.04	1.12	1.23
5	with socket, C6 omitted, *no optimization*	1.07	1.13	1.23	1.28	1.14	1.21	1.33
6	with socket, C6 omitted, C7 optimized at 432 MHz	1.05	1.08	1.13	1.17	1.09	1.23	1.34
	Hewlett-Packard 343A	1.04	1.05	1.08	1.06	1.09	1.14	1.17
	Hewlett-Packard 343A	1.02	1.03	1.05	1.06	1.12	1.18	1.23
	Hewlett-Packard 343A	1.03	1.05	1.07	1.08	1.17	1.16	1.13

ultimate, the inductance of L3 can be varied slightly by stretching or compressing the turns to minimize the VSWR. There will be some interaction between the inductance of L3 and the capacitance of C6, so that one will have to be readjusted if the other is changed. Alternatively, a trimmer capacitor can be substituted for C7 and its capacitance varied, in lieu of adjusting L3, to achieve the same results. Again, there will be interaction with the setting of C6, necessitating alternate adjustments of both capacitors.

It can be seen from the figures in **table 1** that even without VSWR-measuring equipment, careful construction should result in a noise source which is as good as, and probably better than, any which have been previously built (or described) by amateurs.

The final VSWR must be measured with both halves of the enclosure assembled. If the VSWR increases when the cover half is in position, it will be necessary to drill an access hole in it so that the trimmer capacitor can be adjusted with the cover in place.

power supply

The power supply circuit which is used with the

The object of the power supply is to establish a constant diode plate current by controlling the filament current and hence, the filament temperature. To accomplish this, op amp U2 is used to drive Q1, a Motorola MJE801 Darlington pair, which supplies current to the 5722 filament. A reference voltage is established at the inverting input of U2, derived from the regulated 5-volt output of U1 and set by R5, the *diode current* control. The diode plate-current return is through resistor R4 connected to the output of U1; a portion of the voltage drop across the resistor is applied to the noninverting input of the op amp by means of the voltage divider formed by resistors R7 and R8.

Once the diode plate current has been set by means of R5, any change in plate current will result in a change in voltage at the noninverting input of U2. Since the sensing circuit is in the negative return of the power supply, an increase in current will cause the noninverting input to become more negative. The resultant decrease in the output of the op amp will lower the base current of Q1, thereby decreasing the diode filament current to compensate for the increase in plate current. Conversely, if the diode plate current tries to fall, the noninverting input of U2 will become more positive. This will increase the output of U2, increasing the base current of Q1, and raising the diode filament current to negate the decrease in plate current.

The thermal inertia of the diode filament introduces a considerable time lag between the time that an error voltage is detected and the time that the filament current is compensated. Therefore the op amp, instead of operating in a linear mode, functions more like a comparator in that its output switches between a level equal to the voltage on pin 4 and about 1.5 volts less than the voltage on pin 8. The op amp output and the diode filament current are a series of rectangular pulses whose frequency varies between approximately 500 and 1000 Hz and whose duty cycle also varies; both the frequency and duty cycle depend on the setting of the *diode current* pot.

Zener diode CR6, connected between the base of Q1 and the diode filament return, has been incorporated to clamp the base voltage, and hence the diode filament current, to a safe value should there be any failure in the regulating circuit. Since the peak-to-peak amplitude of the voltage at the base of Q1 must be very close to 6 volts in order to obtain 20 to 25 mA diode plate current, a five per cent zener diode has been specified to preclude clamping at too low a voltage.

The positive side of the diode plate supply is returned to ground through a 1-milliamp meter and a selected shunt. Three usable ranges, of 3, 10, and 30 mA, have been incorporated; the maximum diode current is between 20 and 25 mA with the circuit shown. The values of the meter shunts (R12, R13, and R14) will depend on the internal resistance of the meter you actually use. R12 must be equal to the meter resistance divided by two, R13 to the meter resistance divided by nine, and R14 to the meter resistance divided by twenty-nine.

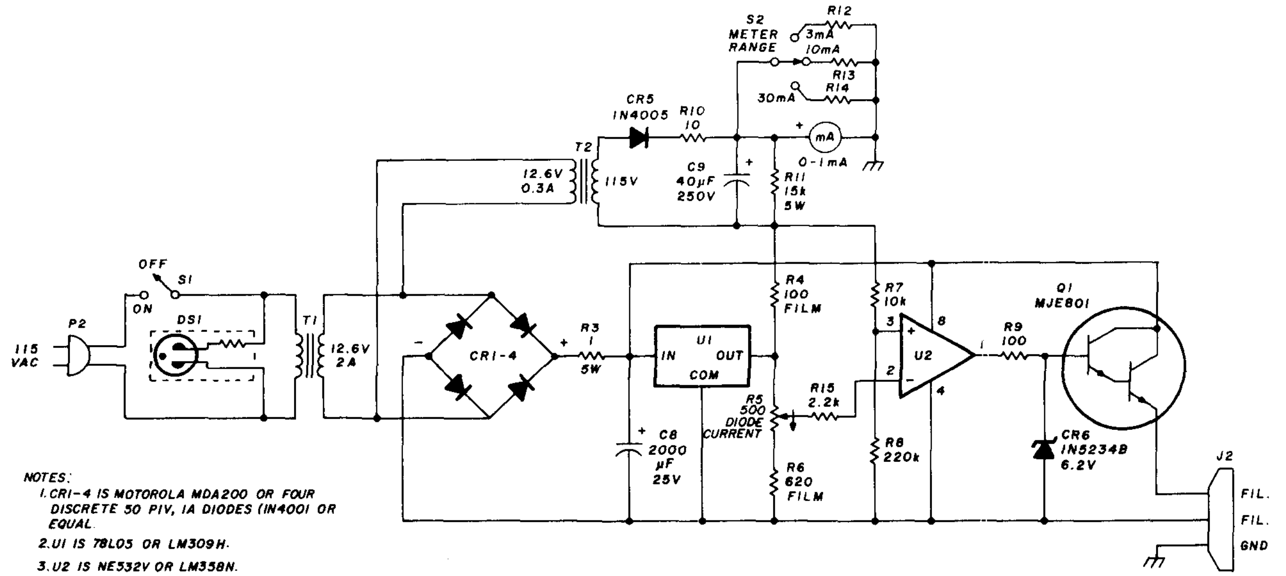

fig. 7. Schematic diagram of the noise-source power supply. The values of R12, R13, and R14 are discussed in the text. Film resistors may be deposited carbon or metal. The *meter range* switch must be a shorting type to protect the meter when the switch is rotated.

power supply construction

The power supply can be housed in any convenient enclosure and can be built in the same way as any conventional power supply. Only a few precautions need be observed. The negative terminal of plate-supply filter capacitor C9 should be returned directly to the transformer winding, and the negative side of the filament-supply filter capacitor C8 should be connected directly to the negative terminal of the bridge rectifier. Otherwise excessive ripple can appear in the outputs.

Transistor Q1 must be mounted on a heatsink which is insulated from chassis ground, or it must be insulated from a grounded heatsink by means of a mica or plastic insulator. If chassis-type construction is used, the chassis can serve as the heatsink. Otherwise, a heatsink having a radiating area of approximately 65 square cm (10 square inches) should be satisfactory. Be sure to apply a thin coating of silicone heatsink compound to the heatsink side of the transistor, and to the insulator if one is used.

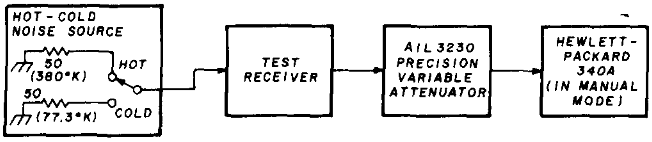

fig. 8. Test setup used to determine the test-receiver noise figure by means of a hot-cold noise source.

As mentioned previously in the discussion of the noise source, *J2 can be any female connector* which will mate with the noise-source cable connector. Inexpensive types, such as the Amphenol 91-MPM3L for the noise-source cable and Amphenol 78-PCG3 for the power supply, are more than adequate.

A word about the meter is in order. Since the accuracy of the noise-figure measurement is a function of the accuracy with which the diode current is measured, a high-quality meter should be used. The accuracy of shunt resistor R12 is equally, if not more, important. (R13 and R14 are of lesser importance, since the accuracy of noise-figure measurements on these ranges is generally not critical.) If possible, the accuracy and tracking of the 3-mA meter range should be checked against a milliammeter of known accuracy.

One way of trimming R12 to the proper value is to use a one per cent metal-film resistor whose resistance is the closest value higher than one-half the meter resistance. Then shunt the metal-film resistor with higher-value composition resistors (starting at about ten times the one per cent value) until the meter reading is correct.

After double checking the circuit, connect the noise source to the power supply. Set the *diode current* control for minimum current and set the *meter range* switch to its *30-mA* position. Energize the power supply and slowly increase the diode plate current by means of the *diode current* control. The current should increase smoothly from zero to a maximum of between 20 and 25 mA. If a plate current of at least 20 mA cannot be reached, it is likely that CR6 is limiting the base voltage of Q1. Reduce the plate current to zero, turn off the power supply, and disconnect one side of CR6. Then turn on the power supply again and increase the diode plate current to its maximum. If the plate current is now higher than could be obtained with CR6 connected, another zener diode should be used which has a slightly higher zener breakdown voltage. Alternatively, the existing zener voltage can be increased by about 0.6 volt by connecting an ordinary silicon diode in series with CR6; the diodes should be connected anode-to-anode.

If current limiting is not being caused by the zener diode, check the various voltages throughout the power supply. They should approximate the values listed below. Except for the output of U1, a 20 per cent variance in voltages can be expected because of component tolerances, especially in the transformers.

	voltage measurement at	
measure across	**1 mA plate current**	**20 mA plate current**
T1 secondary	13.3 Vac	12.4 Vac
T2 secondary	113 Vac	88 Vac
R3	0.9 Vdc	1.5 Vdc
C8	13 Vdc	10 Vdc
U1 output (to common)	5.0 Vdc	5.0 Vdc
CR6	6.2 Vp-p	6.2 Vp-p
V1 filament	4.4 Vp-p	4.7 Vp-p
R11	123 Vdc	72 Vdc

The voltages across CR6 and the filament of V1 must be measured using an oscilloscope because they consist of rectangular pulse trains. Use extreme care when making such measurements, since the scope ground connection will be elevated from the power-supply chassis ground by the plate-supply voltage.

The two characteristics of the noise source which required evaluation were the VSWR and the *ENR*. The former presented no problem, since suitable VSWR-measurement equipment was available. **Table 1** shows the VSWR of six of the homebuilt noise sources at various frequencies between 30 and 550 MHz. For comparison purposes, the measured VSWR of three Hewlett-Packard model 343A noise sources are also included. It can be seen that the comparison is favorable, especially at the amateur band frequencies of 432 MHz and below.

Measuring the excess noise ratio is another matter, however. Short of sending a unit to the National Bureau of Standards, or equipping a primary standards laboratory, there is no way to measure the noise output quantitatively. As derived earlier in this article, the theoretical *ENR* of a temperature-limited diode, when terminated by a resistive load at the standard reference temperature of 290°K, is nearly equal to $10 \log I$ in a 50-ohm system, where I is the diode plate current in milliamperes.

While we might accept this expression as an abso-

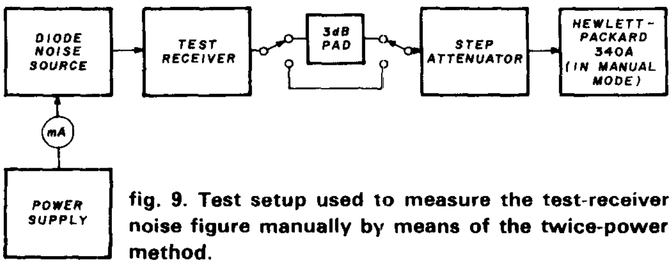

fig. 9. Test setup used to measure the test-receiver noise figure manually by means of the twice-power method.

lute value at frequencies below 30 MHz, we know from experience and from Hewlett-Packard's specifications for the 343A noise source that the *ENR* gradually increases above 30 MHz to the limit of the source's frequency range. (This frequency limitation results from the transit time of the electrons passing from cathode to anode becoming an appreciable part of the period at high frequencies, and from the series inductances shown in **fig. 4**.)

Since there was no method by which the actual noise output of the diode could be measured directly, we resorted to measurement by transfer. While this is a far from ideal technique, it was the only feasible method, and provided us with usable data. The transfer method used was as follows. A hot-cold noise source and a precision variable attenuator were used in conjunction with a Hewlett-Packard noise-figure meter (in its manual mode as an indicator). The test setup is shown in **fig. 8**.

The noise source comprised two 50-ohm terminations, one immersed in liquid nitrogen (boiling point at 77.3°K) and one housed in a thermostatically controlled oven (temperature at 380°K). The VSWR of each termination was measured through the coaxial switch and found to be less than 1.02:1. The noise figures of four receivers, one each at 28, 144, 220, and 432 MHz, were determined by means of the test setup shown, taking into account the measured insertion loss of the coaxial switch (approximately 0.05 dB at 432 MHz, and insignificant at lower frequencies). These noise figures provided the reference for the transfer measurement.

The noise figures of the same receivers were then remeasured, using six different homebuilt noise sources and two Hewlett-Packard 343A noise sources. In each case, the noise figure was determined manually, by means of the twice-power method, and automatically, using a Hewlett-Packard model 340A Automatic Noise Figure Meter; the test setups for these measurements appear in **figs. 9** and **10**.

The results of these noise-figure measurements were analyzed and plotted as errors in the *ENR* of each noise source, compared with the *theoretical ENR* of a temperature-limited diode. The results are shown in **fig. 11**, along with the nominal and specified limits of the Hewlett-Packard model 343A *ENR*

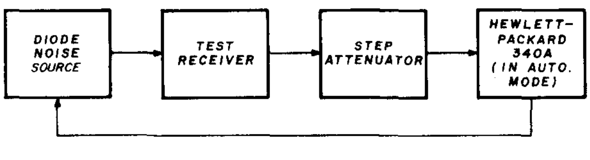

fig. 10. Test setup used to measure the test-receiver noise figure by means of an automatic noise-figure meter.

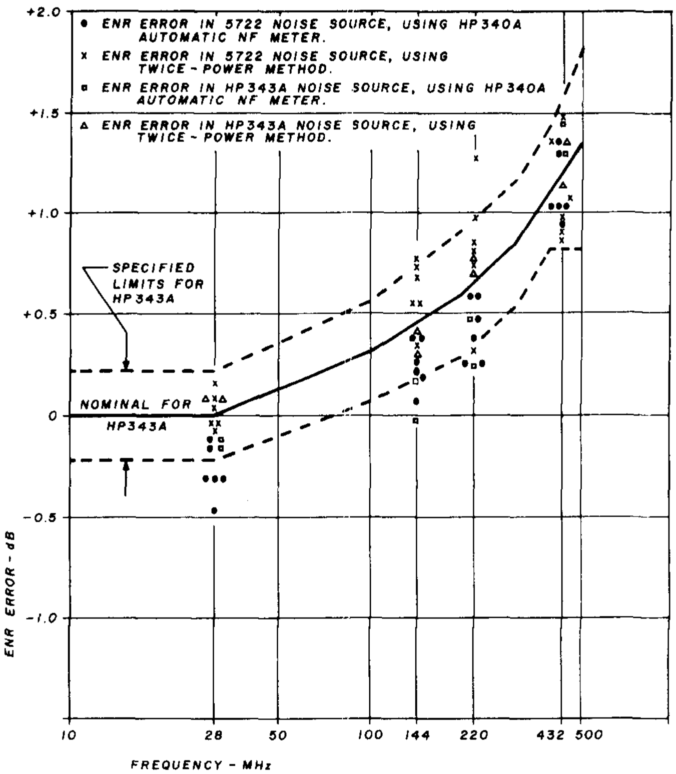

fig. 11. Errors in excess noise ratios of six homebuilt 5722 noise sources and two Hewlett-Packard model 343A noise sources, compared to theoretical noise power. Errors were determined by comparison with receiver noise figures measured using a hot-cold noise source. A positive error indicates output noise power in excess of the theoretical noise power; therefore a measured noise figure must be increased by the indicated error.

error. It must be reiterated that the *ENR* error is the deviation from the theoretical low-frequency *ENR*, and is a normal and expected deviation. Our tests were made to determine the magnitude of the deviation.

Analyzing the results of these measurements proved to be somewhat more difficult than performing the tests. At 28 MHz we expected the *ENR*s to be very close to the theoretical value. This proved to be the case for the manual measurements, but the automatic noise-figure measurements were generally on the high side, indicating a lower *ENR* than expected. The conclusion was that the error was in the automatic noise-figure meter, which is specified as having a possible error of ±0.5 dB. By way of comparison, the milliammeter shown in **fig. 9** was an accurate digital meter, the loss of the 3-dB pad had been checked at 30 MHz, and the noise-figure meter was used only as a reference indicator, so that any error in that instrument was eliminated by using a fixed meter reference reading. The trend of low readings in the automatic mode is continued at 144 and

220 MHz but not at 432 MHz, where both types of measurements yielded closely related results.

Although one of the homebuilt noise sources produced an *ENR* well outside of the expected range at 220 MHz (for some unexplainable reason), the results seem to indicate that a homebuilt noise source can be constructed and can be expected to provide an *ENR* within the range specified by Hewlett-Packard for their model 343A. No attempt was made to apply a temperature or mismatch-loss correction, since the intent was a comparison between the homebuilt noise source and the Hewlett-Packard 343A. Both types of noise sources were subjected to the same possible receiver mismatch and were operated under identical environmental conditions.

noise-figure measurements

As previously stated, the noise source can be used in conjunction with its own power supply to make noise-figure measurements, or it may be used with a Hewlett-Packard model 340B or 342A Automatic Noise Figure Meter. It may also be used with a modified Hewlett-Packard model 340A, as described in reference 1. The techniques employed in automatic noise-figure measurements are treated briefly in that article, and in detail in the Hewlett-Packard manuals covering such equipment.[6-9] The homebuilt noise source, when equipped with the appropriate power connector, can be considered as a direct replacement for the Hewlett-Packard model 343A at frequencies to at least 450 MHz.

Table 2 summarizes, for the amateur frequencies of interest, the nominal noise-figure correction which must be added to the measured noise figure because of the noise-source *ENR* increase with frequency, as shown in **fig. 11**. The table also includes the accuracy of the *ENR*, based on Hewlett-Packard's specifications for the model 343A noise source and the measured equivalence of the homebuilt versions. Thus, this is the uncertainty of the corrected noise figure due to *ENR* only, but does not include other errors discussed in reference 1.

Since the *ENR* is higher, at frequencies over 30 MHz, than the nominal 5.2-dB value for which Hewlett-Packard automatic noise-figure meters are calibrated, it follows that the *ENR* may be reduced to 5.2 dB by reducing the diode plate current. The required value of current plotted against frequency is shown in **fig. 12**, and is included in **table 2**. By setting the diode current to the value indicated for the frequency of measurement, no *ENR* correction need be applied to the noise-figure reading. Manual noise-figure measurements may also be made in conjunction with an automatic noise-figure meter, as described in the previously referenced manuals.

table 2. Nominal corrections for temperature-limited diode noise sources at Amateur frequencies.

frequency (MHz)	*ENR* accuracy (dB)	*NF* correction (dB)	*ENR* (dB) at 3.31 mA	automatic noise-figure meter diode current (mA) to compensate for *NF* correction
28	±0.20	0	5.20	3.31
50	±0.23	+0.13	5.33	3.21
144	±0.28	+0.45	5.65	2.99
220	±0.30	+0.65	5.85	2.85
432	±0.38	+1.17	6.37	2.53

Manual measurements, using the power supply described in this article, can be made in one of two ways: the twice-power method and the Y-factor method. The former is the one most familiar to most amateurs, and will be covered first. **Fig. 13** shows four configurations, in order of preference, for the twice-power noise-figure measurement. In each diagram, the "receiver under test" means any receiver or portion thereof (such as a converter or mixer) which provides an output at either an intermediate or audio frequency. If AGC is incorporated in the receiver, the AGC should be disabled and the rf gain control

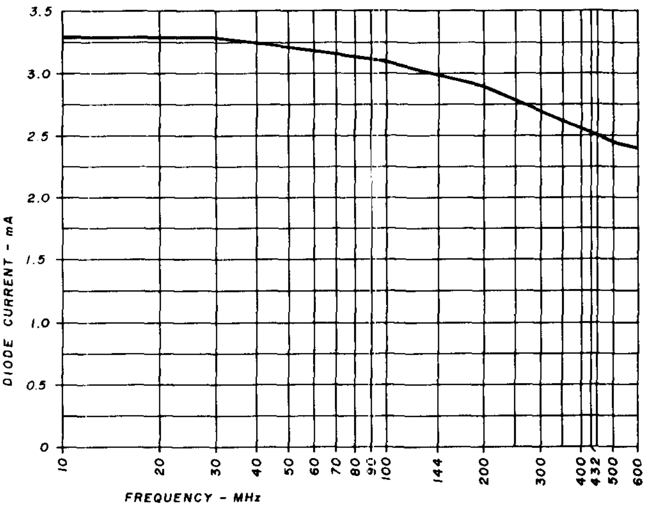

fig. 12. Diode current required by a temperature-limited diode noise source to maintain a nominally constant 5.2-dB excess noise ratio (curve by J. R. Reisert, W1JR).

set so that the receiver is not overloaded by the noise input. There should be a direct connection between the noise-source output connector and the loss pad, and between the pad and the receiver. This means *no cables*, and a minimum of adapters.

The use of a loss pad with a diode noise source is not absolutely essential, but can minimize several problems. First of all, most receivers do not present a 50-ohm input when optimized for a noise match. Because the rated *ENR* of a noise source is based on a 50-ohm load, there will be an indeterminate mismatch loss if the receiver VSWR is greater than 1.0:1. A 3 to 6 dB pad will not eliminate the mismatch loss, but may reduce it somewhat.

A second reason for using a loss pad is to ensure a 50-ohm source impedance for the receiver, since the VSWR of the noise source is not a perfect 1.0:1. Any tendency of the receiver to "take off" when looking into an impedance other than 50 ohms will be reduced by the use of a pad.

In **fig. 13A**, the i-f output of the receiver under test is connected to a video or rf voltmeter (depending on the receiver output frequency), which is used only as an indicator. To terminate the 3-dB pad properly and present a constant load impedance to the receiver, the voltmeter must have a 50-ohm input impedance.*

With the noise source off and the 3-dB pad out of the circuit, a reference reading is established on the voltmeter. The 3-dB pad is then inserted between the receiver and the voltmeter, the noise source is turned on, and the noise-source plate current is adjusted until the same voltmeter reference is obtained. The uncorrected noise figure (in dB) of the receiver plus the loss pad between the noise source and the receiver is equal to $10 \log I$, where I is the diode plate current in milliamperes. The uncorrected receiver noise figure is determined by subtracting the attenuation of the loss pad from the calculated noise figure. The receiver noise figure must then be corrected for frequency by adding the noise-figure correction listed in **table 2** or, at frequencies not listed in the table, the nominal Hewlett-Packard 343A error shown in **fig. 11**. Note that the noise-figure accuracy is limited by the degree of uncertainty in the noise-source *ENR*, as indicated in **table 2** and **fig. 11**. Other uncertainties in the measurement will be discussed later.

The advantages of the circuit shown in **fig. 13A** are twofold: the measurement is independent of both the voltmeter calibration and any nonlinearity in the receiver detector. **Fig. 13B** shows a similar setup, except that the 3-dB increase in noise power depends on the voltmeter calibration. In this arrangement, the voltmeter ideally should be a true rms type so its readings are proportional to the square root of the power. It need not have 50-ohm input impedance, but should be calibrated in dB, although the latter feature is not absolutely essential. An average-responding voltmeter can be used with little loss in accuracy, but a peak-responding type should be avoided if at all possible.†

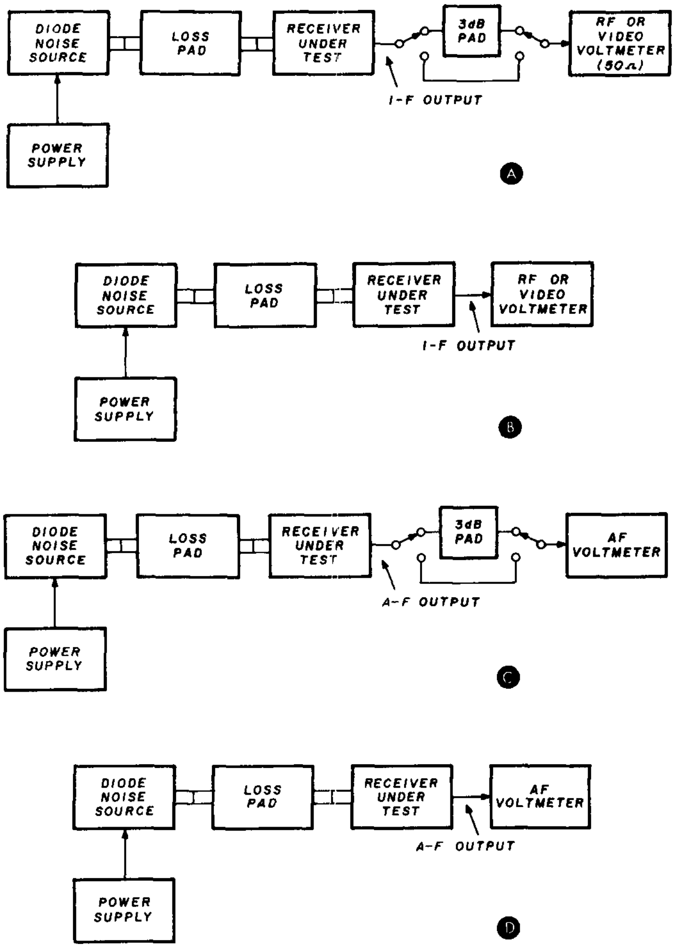

fig. 13. Four test configurations for measuring noise figure by the twice-power method, shown in order of preference. In A and C, the measurement is independent of the voltmeter calibration. In A and B, the measurement is independent of the linearity of the receiver detector.

*A high-impedance meter can be converted to 50 ohms by means of a 50-ohm, feed-through termination, such as the Heath SU-511-50, Tektronix 011-0049-01, Hewlett-Packard 10100C, or Systron-Donner 454.

†The terms "average-responding" and "peak-responding" refer to the voltmeter circuit, not the meter scale calibration. The Hewlett-Packard model 400D is a typical average-responding meter calibrated in rms volts. Voltmeters which employ rf probes, such as the Hewlett-Packard model 410B and the various Heath, Eico, and similar electronic voltmeters, are invariably peak-responding meters, with their meter scales also calibrated in rms volts. In all of these instruments, the rms meter calibration is based on a sinusoidal waveform, which is not valid for noise voltage.

When the arrangement of **fig. 13B** is used, a reference reading is established on the voltmeter with the noise source off. Then the noise source is turned on and the diode plate current is adjusted until the meter reading is exactly 3 dB greater than the reference reading. (If the voltmeter does not have a dB scale, the diode current should be adjusted until the meter reading is exactly 1.41 times the reference voltage.) The noise figure is determined as described for **fig. 13A**.

Figs. 13C and **13D** correspond to the test circuits just discussed except that the audio output of the receiver is measured by means of an audio-frequency voltmeter. Both of these circuit configurations depend on the linearity of the receiver detector; fortunately, most modern receivers use product detectors which are generally quite linear over a 3-dB range. The receiver beat-frequency oscillator must be on, of course, for the product detector to function.

Noise-figure measurements are made exactly as described for **figs. 13A** and **13B**. In **fig. 13C**, the impedance of the voltmeter need not be 50 ohms, but its impedance and that of the 3-dB pad must be the same. In some cases, it may not be convenient or even possible to disable the receiver AGC when making measurements in accordance with **figs. 13C** or **13D**. In view of the fact that the noise levels introduced are extremely low, the AGC in most, if not all, receivers will not be activated; so there is very little likelihood of any error occurring if the AGC cannot be disabled.

The less familiar Y-factor method of measuring noise figure requires a precision variable attenuator, as shown in **fig. 14**. The resolution of the attenuator should be at least 0.1 dB, and preferably 0.01 dB, and must be of known accuracy at the converter output frequency. With the noise source off, the precision attenuator is adjusted to obtain a convenient reference reading on the voltmeter. The noise-source plate current is then set to a value which corresponds

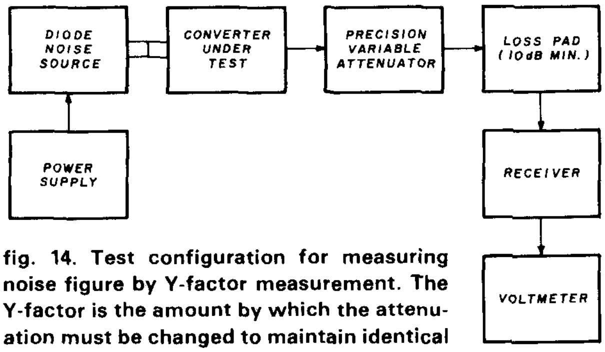

fig. 14. Test configuration for measuring noise figure by Y-factor measurement. The Y-factor is the amount by which the attenuation must be changed to maintain identical meter readings with the noise source off and on.

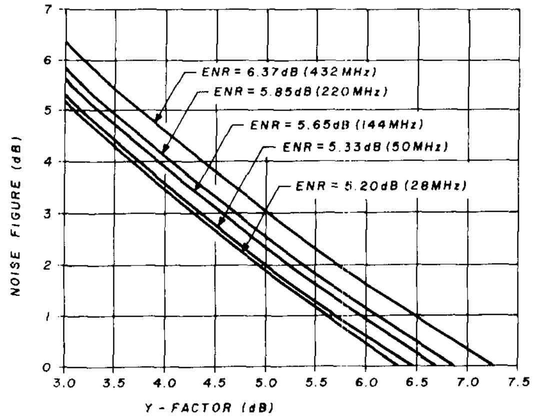

fig. 15. Noise figure plotted as a function of Y-factor for the nominal excess noise ratios of a temperature-limited diode operating at 3.31 mA plate current.

to a known *ENR*, *e.g.*, 3.31 mA, and the precision attenuator is readjusted to obtain the reference meter indication. The difference between the two attenuator settings, when converted from dB to the equivalent power ratio, is called the Y-factor, and is related to noise figure by the expression

$$NF = ENR - 10 \log (Y - 1) \qquad \textbf{(18)}$$

where both the noise figure (*NF*) and excess noise ratio (*ENR*) are in dB.* **Fig. 15** shows noise figure plotted against Y-factor (expressed in dB) for the nominal excess noise ratios obtained when the diode current is set to 3.31 mA. These curves obviously eliminate corrections because of differences in *ENR* at various frequencies, but are still subject to the uncertainty of the *ENR* tolerance.

There is a third method of manually measuring noise figure which is not recommended, but which should be mentioned because of its past appearance in some publications. It entails inserting a precision variable attenuator between the noise source and the receiver under test, then determining the noise figure from the *ENR* of the source and the calibration of the attenuator. This method is generally inaccurate because of variations in attenuator accuracy with frequency, and because the load impedance presented to the noise source by the attenuator-receiver combination may change as the attenuator setting is changed.

measurement errors

Some of the possible sources of error in noise-figure measurements are already apparent from the preceding discussion, especially that which is inherent in the noise source itself. However, there are addition-

*This relationship is derived in **eqs. 1** through **9** of reference 1.

al errors which must be considered and which are discussed in greater detail in reference 1. These errors comprise the following:

1. Noise-source accuracy, corrected for frequency

2. Noise-figure meter or power-supply milliammeter accuracy

3. Receiver image-response error

4. Temperature error

5. Mismatch error

If all errors are in dB, they accumulate additively. Therefore, the total measurement error will be the sum of the above. This is an imposing list, and could total well over 1.5 dB if all errors were of the same algebraic sign. However, many of these errors will cancel because of opposing signs, and generally the accuracy of commercial test equipment is better than the limits of its specifications. Nevertheless, these possibilities of error are very real and cannot be ignored except for *comparative* measurements using the same equipment at one particular time. And even then, the mismatch errors between the noise source and different receivers under test still exist.

acknowledgments

Bob Melvin, W6VSV, was responsible for both the noise-source analysis and the power-supply design, and must share credit for this article. Bryan Westfall, K6OJM, and Steve Mieth, K6YFK, provided equipment and invaluable assistance in making the hot-cold measurements used to evaluate the performance of the noise source. Appreciation is also due Cliff Buttschardt, W6HDO; Duke Moran, W6SPB; Paul Shuch, N6TX; and Bob Sutherland, W6PO; for the use of their commercial and homebuilt noise sources for the evaluation process.

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ground currents measuring in 160-meter antenna systems

Several years of operating on 160 meters has verified one thing for sure: The antenna is the name of the game. All the store-bought boxes and smooth-turning dials are useless without that super skyhook. This article describes an instrument that will tell you the behavior of rf ground currents in your antenna system: where it's going, how much, where the current divides, and the location of any conducting element within the instrument's field.

top-band antennas

Few Amateurs are lucky enough to have room for even a simple dipole on 160 meters. Such an antenna requires a length of at least 76 meters (250 feet) and it won't work too well as a DX antenna unless it's 30 meters (100 feet) or more in the air.

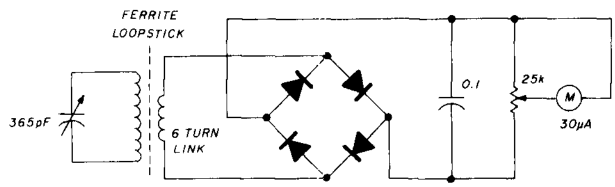
Loading a tower or short vertical antenna is often the best that space will allow. Many articles have been published on how to load this or that antenna on 160 meters and make it work. I've tried many and found that most have a common problem: *ground losses*. The shortest piece of wire that will resonate is ½-wavelength long. That's 78 meters (257 feet) at 1824 kHz.

Shortened vertical antennas will work, but all verticals shorter than ½ wavelength must use the earth as a mirror image to supply the missing portion of the antenna. A typical example is the ¼-wavelength vertical that works against its mirror image to produce the patterns you see in the *Radio Amateur's Handbook*. The problem is usually found in the not-so-perfect ground around your station. In most cases this lossy ground becomes part of your antenna system. This is particularly true for short verticals and inverted L antennas. These antennas are electrically in series with the lossy ground. Working with such an antenna can be frustrating; often you can hear well enough, but much of your transmitter power ends up heating the worms.

Many of the better signals on 160 meters come from hams that have buried literally thousands of feet of copper in the ground to reduce these losses. The prospect of putting that much wire into the ground makes my back ache just thinking about it. Can't you hook onto your water pipes for some of that ground? How about the backyard fence? Would it help to drive a few ground stakes? All the mathematical gymnastics I could muster just didn't tell me where those ground currents actually went. What I needed was a device that would measure the rf current in the ground system and tell me where it was actually going. I finally built a device that does just that: it's called a magnetometer.

the magnetometer

The idea of a magnetic fieldstrength indicator is far from new. An article in the older *ARRL Antenna Handbooks* describes a less-sensitive model for higher frequencies. The 160-meter instrument described here uses a six-turn link and full-wave rectifier to make the impedance match to the meter more effective. Somehow the unit works better on 160 meters than on the higher-frequency bands. One reason is that it's harder to get a proper ground on 160 meters, and ground currents tend to be higher for longer distances through every possible path. By using the magnetometer you can tell the relative current and direction in *any* wire near your antenna or ground system — or in your antenna itself, for that matter. You can locate a buried ground or radial by sweeping



quantity	description
1	aluminum minibox, 102 × 128 × 77 mm (4 × 5 × 3 inches)
1	25-kilohm potentiometer
1	0-30 or 0-50 microammeter
4	germanium diodes (1N34 or equivalent)
1	102-mm (4-inch) broadcast ferrite loopstick (about 6.5 mm or ¼ inch — round works best)
1	365-pF variable
1	0.1 μF disc ceramic
1	terminal strip
2	rubber grommets

fig. 1. Schematic of the homebrew magnetometer.

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the magnetometer over the area and watching the meter deflection. You can compare relative currents in different radials or grounds.

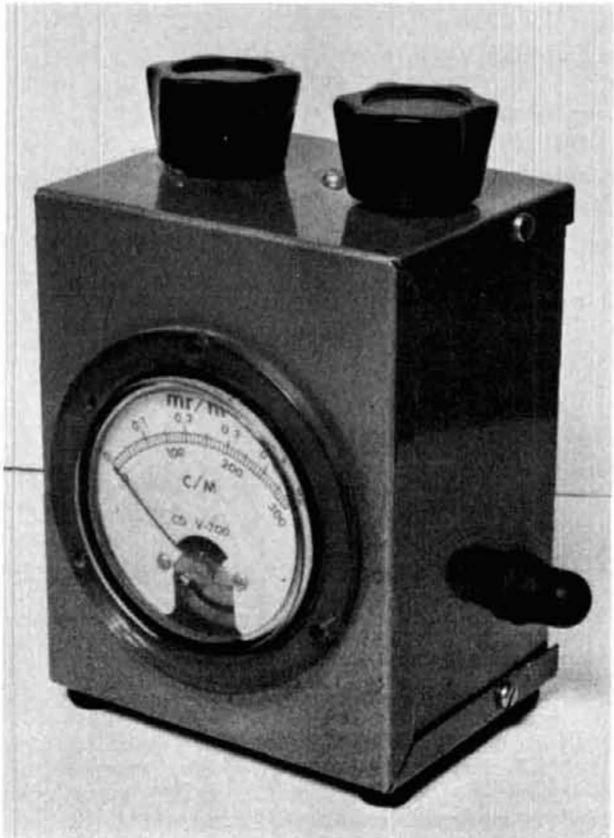
In my basement, where the rf current in the water pipes comes to a T intersection, I can measure how the current divides, if it does, and the relative currents in each section of pipe. In my case, the gas pipe was a better ground than the water pipe!

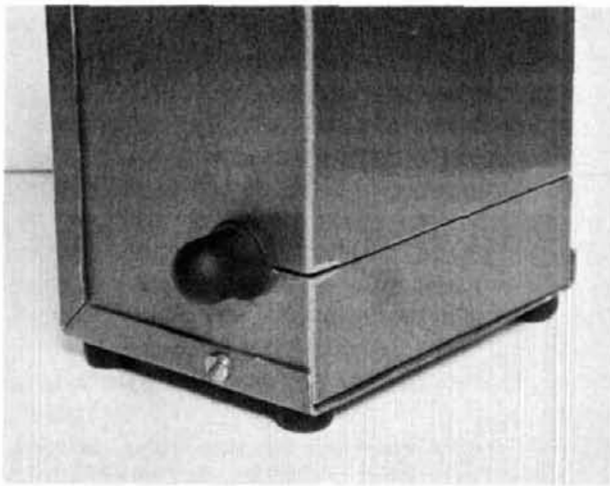
The magnetometer is not sensitive to the antenna's radiated field. It is shielded from this field by the aluminum box and will not act as a field-strength meter unless held perpendicular to a conductor carrying rf current. The slot must be aligned perpendicular to a conductor to allow the magnetic field to reach the loopstick.

It is this magnetic field that operates the device. Fifty watts of rf power into your antenna is usually enough to give a useful reading. (Do your testing during the daylight hours when 160 meters is quiet.) A few minutes with the magnetometer will tell you exactly where your rf energy is flowing.

Unless your antenna is textbook perfect, and few are, you're sure to get a few surprises. I checked WØNFL's station in this manner. Jim was using an inverted L antenna and loaded it through a series-tuned capacitor. Although he had several ground stakes and many buried ground wires, some rf still

Front view of the magnetometer. Small, inexpensive, and easy to build, this device traces one of the most common problems on 160 meters: ground losses.





Rear view of the magnetometer. The slot must be aligned perpendicular to a conductor to allow the magnetic field to reach the loopstick (see fig. 1).

flowed into the station. All pipes and the furnace air duct in the basement were bonded together and tied to the ground system.

I still remember Jim's surprise when we found that he was loading his kitchen sink. It turned out that the soil pipe provided the best rf ground, and the only path to that was through the S-trap under the kitchen sink. Boy was it hot!

I've used my magnetometer for about four years and have had nothing but good luck and fun with it. I sent a sketch to a couple of my 160-meter cronies, W9GDW and NØBD. Both report that it works well for them too.

construction

The best part about the magnetometer is that it is inexpensive, requires no power supply, and most hams can build it in one evening, often from parts in the junk box. Parts placement isn't critical, and I'm sure that any reasonable approximation of the schematic (**fig. 1**) will work. Use the most sensitive meter you can find and germanium diodes for best results. The loopstick I used was made to operate in the broadcast band with a 365-pF capacitor. I removed about ten turns from the coil to obtain resonance in the 160-meter band. A grid dipper works well to check tuning range. The loopstick is mounted 25.5 mm (1 inch) from the bottom of the minibox and about 19 mm (¾ inch) in from the back. Drill a 12.5-mm (½-inch) hole in each side of the box at these points. Use a hacksaw to cut a slot from the back of the box down to the holes. I used rubber grommets to mount the loopstick. The pickup winding is a six-turn loop of hookup wire wound over the center of the loopstick winding. A few more holes and there you have it: another gadget that no serious 160-meter antenna nut can live without.

ham radio

More Details? CHECK—OFF Page 110

the digiscope

Eliminate a lot of expensive test equipment by using the digiscope, a self-contained test instrument for TTL circuitry

If you experiment with or repair digital circuitry, this article is for you. Even if you don't, consider the fact that most new equipment contains at least some logic circuitry. Whether you're a digital expert or merely contemplating an upgrade of your trouble-shooting skills, you'll find the digiscope very useful.

To understand my claim, look at a few digital basics. Logic circuitry uses only two levels of voltage, and at times a dc voltmeter is all you need to find a problem or debug an experiment. Unfortunately, voltage levels in many circuits don't stand still, and rapid level changes make the voltmeter almost useless. Professional technicians depend on the oscilloscope to see what is happening in such circuits. The cost of a good oscilloscope may exceed the cost of the device you're trying to check. That's where the digiscope fits in. The objective is to provide oscilloscope function at multimeter prices.

You might be interested in how the digiscope evolved. Last year, several of our club members decided to built the frequency counter described by Jim Pollock.[1] Naturally, some of the counters (including mine) didn't work on the first try. I had worked with digital circuitry previously and enjoyed helping with a few of the counters using an oscilloscope for the repairs. As I used the scope, I began to realize that, for these types of circuits, oscilloscope functions could be approximated using inexpensive TTL devices. The result is that many more people could work with complex circuits without making a major investment in test equipment.

oscilloscope functions

To see how I have approximated an oscilloscope function, take a brief look at the digiscope features.

1. A level indicator serves as a basic voltmeter by indicating a 0 or 1 level in the circuit under test.

2. Pulse duration from 0.1 microsecond to 999 milliseconds can be measured. A variation of this feature allows the measurement of time between the trailing edge of a pulse and the leading edge of the next.

3. The length of time between the leading edges of two pulses at different circuit points can be measured.

4. You can defer the measurement of a pulse until after the occurrence of a pulse at another circuit point. This is similar to an oscilloscope's external sync. In addition, this measurement can be delayed for up to 99 milliseconds after the occurrence of the sync pulse, with a delay resolution of 0.1 millisecond. This approximates the oscilloscope's delayed sweep.

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5. Auxiliary gates are available on the front panel, including a latch to check for pulse occurrence or coincidence between two or three pulses.

6. An event counter, inherent in the design, allows counting up to 999 events with overflow indication. Since some readers may not be familiar with the equivalent oscilloscope functions, I will include examples of digiscope use in circuit explanations.

clock and display circuits

Before looking at uses for the digiscope, I'll examine the clock and display circuits, since they are the key to understanding basic digiscope functions. At the heart of the display circuitry is the 74143 TTL IC which combines the functions of the 7490 and 7447, serving as a decade counter as well as a decoder for the displays. Three of the counters are used, permitting up to 999 counts with a latched overflow indicator. Input to the first 74143 is from the clock, with options of 0.1 microsecond, 1 microsecond, 0.1 millisecond, and 1 millisecond.

The pulse to be measured acts as a gate for the counter, allowing clock pulses to be counted during the time the pulse being measured remains at a logic 1 level. Initially, the counter is reset from the front panel. The next pulse to arrive allows counting. After the first ends, further counting is inhibited so that the display may be read. Input clock pulse options allow a measured range from 100 nanoseconds to 999 milliseconds. Resolution varies from 0.1 microsecond to 1 millisecond. A decimal point is displayed when necessary to indicate either 0.1 microsecond or 0.1 millisecond resolution.

The clock circuit and divider chain is almost identical to that used in the WB2DFA 50-MHz counter[1] referred to earlier. The only change is that a 10-MHz crystal is used to obtain 0.1-microsecond measurement resolution. Although oscillators of this type are usually recommended for use up to 1 MHz, the 10-MHz crystal has worked well. No starting or stability problems have been observed. Counting range is changed by selecting the required frequency via S1A.

As an example of operation, consider the pulses shown in **fig. 1**. Suppose you would like to examine the pulses of **fig. 1B**. To do this, connect the data-in test lead to the appropriate point, and set the range switch to 0.1 millisecond. Push the reset button to enable the scope. During the pulse time, pin 13 of U1 (see **fig. 2**) is a 1 for its duration, or 10 ms. During this 10 ms, one hundred pulses from pin 12 of U4 are passed to the counter/display circuits. At the end of the 10 ms pulse time, the display is then held with 100 displayed. S1B selects the necessary decimal point line causing the display to read 10.0 milliseconds, the length of the pulse. So that I may view the measurement, subsequent pulses are blocked, until the reset is pressed, so that the display may be read.

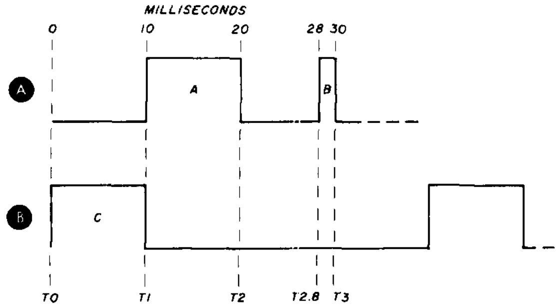


fig. 1. Timing diagram of the two pulses used for explanation of the digiscope's circuits. The time scale is 2 ms per division.

Having measured the pulse width of **fig. 1B**, you might be interested in the length of time between two consecutive pulses. To determine this, simply change S10 to the invert position and press reset. **Fig. 1B** would now appear as a train of 30-ms pulses with 10-ms separation. Using the same circuitry, the scope would now indicate a pulse length of 30.0 milliseconds. Other factors, such as rise and fall times as well as the existence of jitter, have not been determined, however repeated measurements may help determine the extent of any jitter.

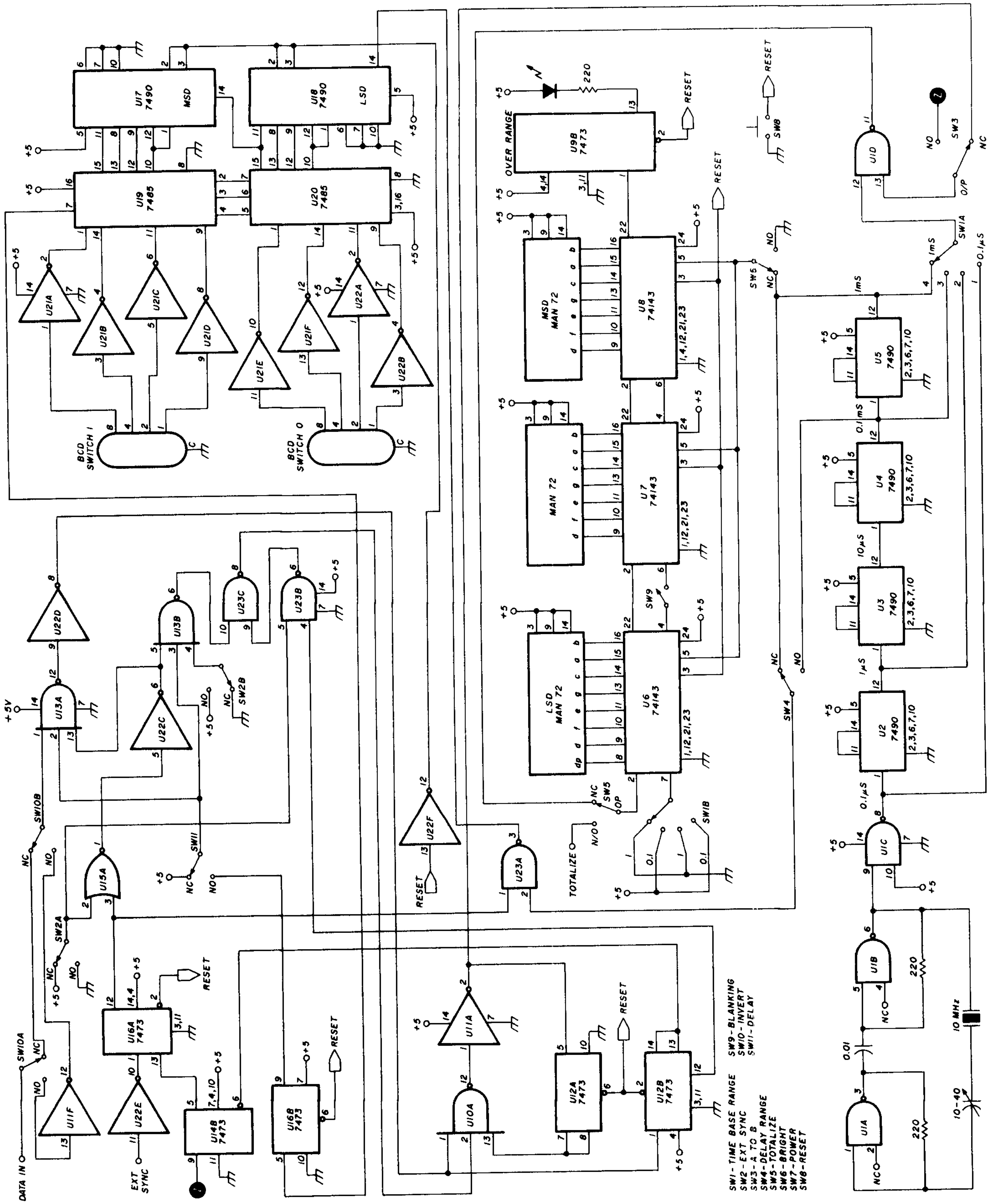
one-shot/swallow circuitry

Recall I said that after the measured pulse ends, further pulses would be ignored. This is accomplished with U12A. The trailing edge of the measured pulse causes U12A to be set, pulling pin 13 of U10 low which prevents subsequent pulses from passing through U10A. This condition exists until the reset switch is pressed.

An interesting problem could arise when the scope is reset. Suppose that the instant after the reset line is grounded you're halfway through a pulse. An erroneous measurement would be obtained. To prevent this, the scope discards the first pulse after reset, ensuring that a partial pulse is not used. After reset, pin 2 of U10 is low, preventing pulses from passing to the counters. The first pulse, whether partial or complete, is blocked with its subsequent trailing edge used to set U12B. Indirectly, this causes pin 2 of U10 to go high. The second pulse is now successfully passed through U10A to the counters. By swallowing the first pulse, I ensure that only complete pulses are measured.

external sync

Examining pulses of uniform length is a fairly easy task. Unfortunately, the pulses may be of varying lengths, as illustrated in **fig. 1A**. By using the gating


SW1-TIME BASE RANGE
SW2-EXT SYNC
SW3-A TO B
SW4-DELAY RANGE
SW5-TOTALIZE
SW6-BRIGHT
SW7-POWER
SW8-RESET
SW9-BLANKING
SW10-INVERT
SW11-DELAY

method I have described, repeated measurements would randomly show pulses of 10 milliseconds and 2 milliseconds. In our simple case, merely finding both pulses may be sufficient. A more precise method is to look for a pulse only at the time it should occur.

As an example, suppose I want to measure the first pulse of **fig. 1A**. Note that this occurs after the pulse of **fig. 1B**. By connecting the external sync input to the appropriate spot and placing the norm/external sync switch in the external sync position, U16A is set by pulse C. The next pulse to arrive at data-in is pulse A, the one I want to measure. In the external sync mode, the pulse-swallowing circuitry is disabled by U13B. Since I know exactly when to expect pulse A, partial pulses are not a problem.

delayed sync

To illustrate the delayed sync logic, assume that you want to look at only pulse B in **fig. 1A**. To avoid random measurement, use the external sync feature. If pulse C began between T1 and T2.7, the external sync feature would be sufficient. However, what is needed is a way to use pulse C as the sync but delay the input until after pulse A.

This cabability is provided by the delayed sync cir-

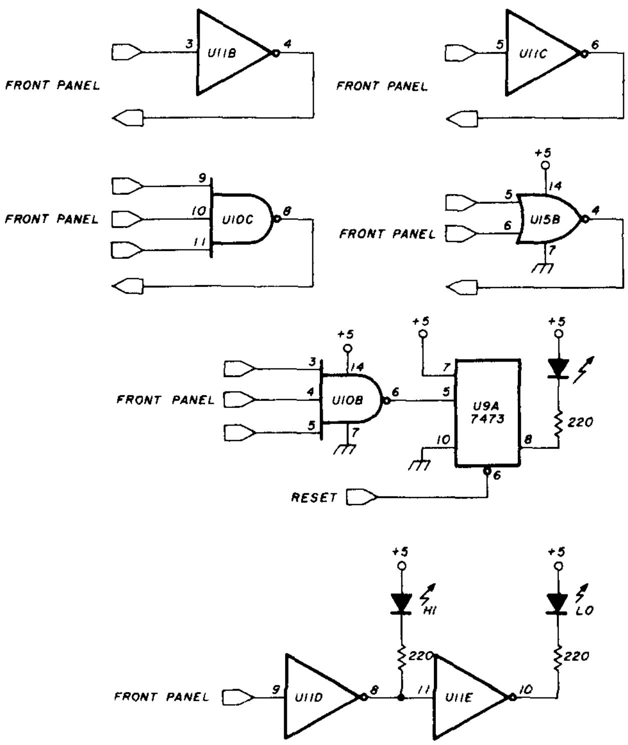


fig. 2. At left, schematic diagram of the digiscope. The RESET switch simultaneously controls all reset functions. Above, the external inputs and conditioning circuits.

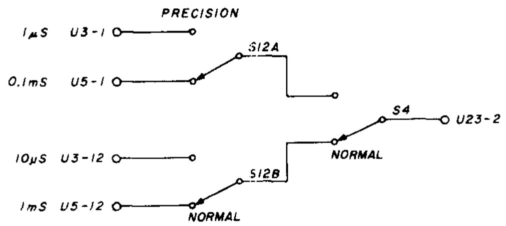


fig. 3. Optional circuit to enable the operator to select smaller increments of delay time.

cuitry. With the digiscope set up as in the last example, place S11 in delay position. Note that in the delay mode, setting U16A is still not sufficient for U13A to pass a signal; U16B must now also be set. This will occur at T0 plus some amount of delay selected on the front panel. In this example, the delay switches are set for 20 ms. Thus, at T2, U16B will set, enabling pulse B to pass through U13A to the counter. In this example, pulse B could be selected with a delay ranging from 11 to 27 milliseconds.

The delayed sync logic allows me to enable measurement at a precise instant after the occurence of the external sync. This delay can range from 0.1 to 99 ms. S4 selects delay increments of 0.1 or 1 ms. Counters U17 and U18 are responsible for delay timing and are held at zero after the reset is pressed. When the external sync pulse arrives at U16A, U17 and U18 are allowed to begin accumulating delay pulses. The desired delay is set into the thumbwheel switches. These switches present a BCD format to U19 and U20, where they are compared with the count in U17 and U18. When the count exceeds the amount set in the switches, a negative pulse is passed to U16B. This results in the set of U16B and the enabling of data-in.

The precision of the sync delay circuit is very good, since it is based on the crystal-controlled clock. Accuracy however, is dependent on the resolution inherent in the delay range you have selected. Actually, the delay begins not at arrival of the external sync, but at the next clock pulse after external sync. Thus delay error could be from zero up to one unit of resolution. Any error always results in early expiration of delay. Suppose I select a 0.1-ms delay increment and set a delay of 4.6 ms in the thumbwheels. Actual delay will vary between 4.5 and 4.6 ms. For greater resolution, I could simply use smaller units of time in incrementing the delay counters. A circuit illustrating optional resolution of 1 or 10 microseconds is illustrated in **fig. 3**. It should be recognized however, that an increase in resolution reduces the total delay available.

Occasionally it may be necessary to measure the time between two pulses at different pins. As an example, I might wish to measure the time between the rise of pulse C and the rise of pulse A. To do this I would connect the pulse A line to the data-in jack, while the other point would be connected to the external sync input and S3 placed in the A-B position. S2 should be placed in the external sync position, S10 in the invert position, and the reset button pressed. When pulse C arrives, it will set U16A, which will set U14B. The Q output of U14B enables the counter input and timing begins. When pulse A arrives at data-in, U12B is set. The $\overline{Q}$ output of U12B resets U14B, ending the counting cycle. With the range switch set to 1 ms, 10 ms will now be displayed.

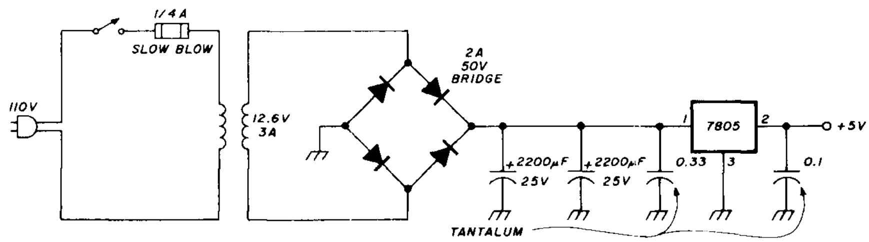


fig. 4. Schematic diagram of a suggested power supply for use with the digiscope.

accessory circuits

To assist in examining unusual situations, some accessory circuits are provided. These are simply logic gates with inputs and outputs extended to the front panel. U11D and U11E provide a basic logic tester. Connecting the tester input to another logic circuit will cause the appropriate front panel LED to indicate a 0 or 1 condition. Note that a 1 will be indicated if connected to an open circuit. Both 0 and 1 will be lit if connected to a pulsing circuit or one with a faulty voltage level between 0 and 1. Each LED is mounted in a 14-pin socket next to the counter display sockets. LED function is indicated by transparent lettering on a negative film placed between the bezel and the LED. The overrange and test latch indicators are also located in this socket.

Also available on the front panel are AND, OR, and inverter circuits. These can be used to combine pins of a circuit for making tests not otherwise possible. An example would be measurement of the overlap of pulses A and C. I could connect the lines of **fig. 1** to the front panel connections of U10. The output could be inverted using U11B. The output of U11B is then connected to data-in. If overlap exists, it can now be measured to the nearest 100 ns.

If I simply wanted to detect the existence of overlap, a test latch circuit is available. It is composed of U10B and U9A. In this example, I could connect to any two inputs of U10B. Overlap would be indicated by the lighting of the test latch display at the output of U9A. This latch is reset with the reset switch.

A totalize function is provided by S5. With this switch in the TOTALIZE position, front panel input is provided to the counters. External events, which are available as TTL pulses, can then be accumulated. I suggest using an accessory inverter to shield the more expensive counter chip from possible excessive voltage levels. Accumulation can be reset at any time with the front panel reset switch.

construction

While building the digiscope I found it difficult to stop adding features. If you build this project, I think you'll find it a good idea to leave room for additions. For example, I think a good addition would be selectable CMOS inputs for working on CMOS circuits. To facilitate this, I left some blank IC positions on my circuit board. While a circuit board is suggested, you could wire wrap this project; my original effort was done on plugboard and didn't seem to suffer too much from all the stray wire.* One note on the 10-MHz crystal is that not any crystal will work well in this circuit. I achieved the best results with a crystal obtained from Jan Crystals. A suggested power supply is illustrated in **fig. 4**.

conclusions

The digiscope is an exciting approach to digital trouble shooting. It can give a way of working on complex circuits without expensive test equipment. If you are interested in digital circuits but don't know how to start, I suggest obtaining the *TTL Cookbook*. I know you'll find working with logic a lot easier than you expected. Whether you're new to TTL or a veteran in logic circuits, I think the digiscope will be a valuable addition to your workbench.

reference

1. Jim Pollock, WB2DFA, "Six-Digit 50-MHz Frequency Counter," *ham radio*, January, 1976, page 18.

ham radio

*A drilled and plated set of circuit boards with additional instructions are available for $19.00 postpaid from RTC Electronics, Box 2514, Lincoln, Nebraska 68502.

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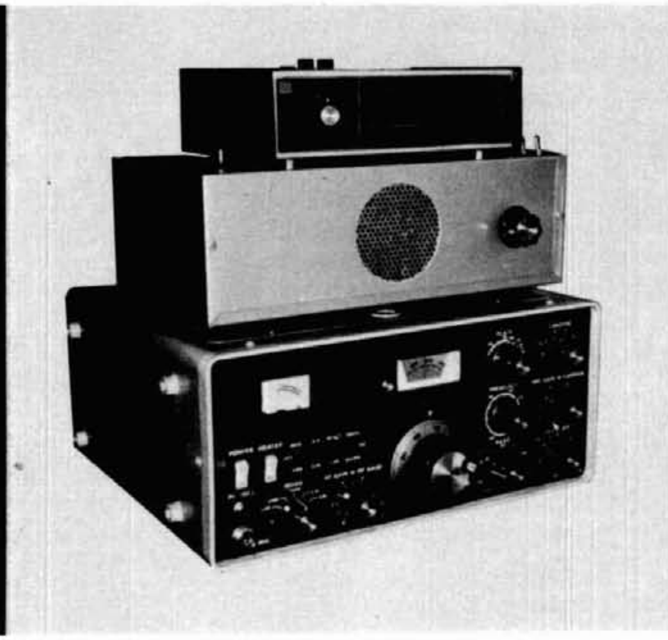


talking digital readout for amateur transceivers

Adapting the talking calculator's synthesized voice to digital displays for the visually handicapped

During a ragchew with Ted Albrecht, W9IFJ, the subject of the talking calculator was brought up. This machine is manufactured by Telesensory Systems in Palo Alto, California, for the visually handicapped. We agreed that it would be most interesting if the talking calculator's synthesized voice were available separately. Perhaps it could be interfaced with digital-readout equipment in general and Ted's in particular.

Happily, upon contacting Telesensory in Palo Alto, we found that the voice synthesizer module was available separately to hobbyists, and its numeric vocabulary was in standard TTL BCD code that already existed at the input of the 7447s in Ted's DD-1 Digital Display Unit.

Ted's purchase of the English-speaking S2A module from Telesensory Systems* represented the committment necessary to get the project moving. Design of an interface between it and the DD-1 is the project this article describes. Much of the design should be credited to Tim Blank, WB9GYU, who is more knowledgeable of digital logic design than I, and who provided the assistance in debugging the assembled unit.

intefacing considerations

Interfacing the voice with the six-digit readout involves changing the parallel (all-at-once) visual display to a serial or sequential form, so that the digits from left to right are said in order with a minimum of delay between words. It would be possible to clock across at regular time intervals if allowance for the longest word were made, but much wasted time would occur, so the clock approach was abandoned. Since each digit can be 0 to 9, or anything in between, four data lines of binary for each are required. To read the digits in sequence then requires that the first four digit data lines of the speech synthesizer module be sequentially connected to the four digit data lines coming from each digit of the visual display. This adds up to twenty-four data lines between the DD-1 and the interface board!

design

The most reasonable approach was to use four eight-position multiplexers, one for each data line. They could be stepped by a binary address of three bits, from 0 through 7. The device that provides this address to the four 74151 multiplexers is a 74193, a most remarkable MSI chip (**fig. 1**). The 74193 can provide a four-line address with a capability of 0 through 15, which would seem to be a problem. However, the 74193 can be reset after any given number of steps very readily. In fact, it can be told to skip a preset number of the initial steps in its output address. Its fan-out is about twenty, indicating no trouble in driving the multiplexers, so it seemed a reasonable basis for an interface. Since the 74193 could be told which addresses to provide, considerable imagination can be used in this aspect of the design.

I decided to use the eight-position capability of the

By Ray Brandt, N9KV, 824 North Marion Avenue, Janesville, Wisconsin 53545

*The S2A Speech Synthesize Module is available from Telesensory Systems, Inc., P.O. Box 10099, Palo Alto, California 94304. Contact Mr. Wladimir N. Walko, Speech Products Manager, (415) 493-2626. Unit price to hobbyists is $95.00.

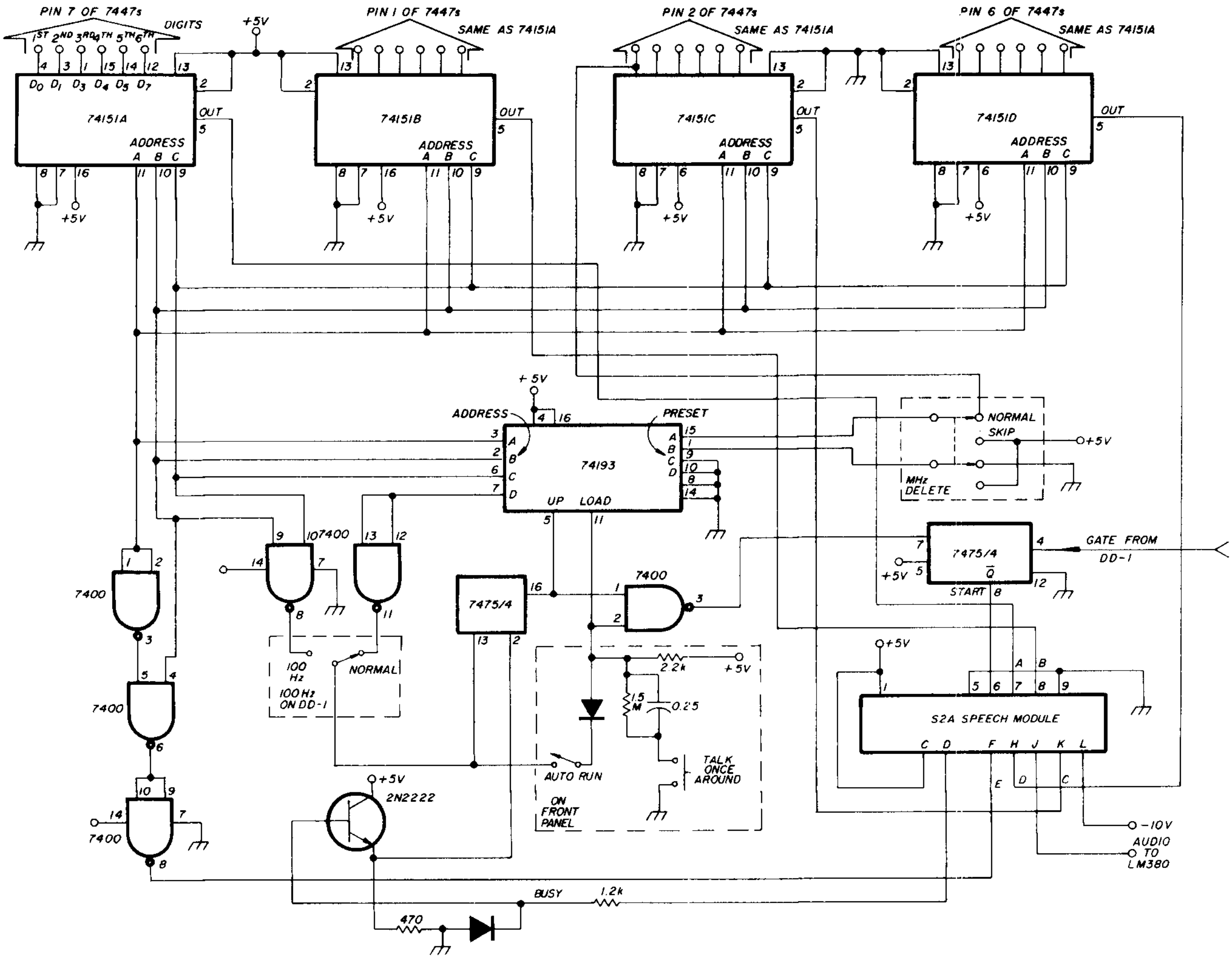


fig. 1. Schematic of the talking readout. The 74193 provides the address to the four 74151 multiplexers. The inputs to the 74151s come from the designated pins of the 7447s in the frequency display.

multiplexers, using the two spare positions to have the voice say POINT between megahertz and kilohertz, and between kilohertz and hertz, provided by the readout unit. Looking at the S2A vocabulary (**table 1**), it's necessary for the **A**, **B**, and **E** data lines to be high, so the **A** and **B** data lines were hard-wired in the 2 and 6 data positions. An additional circuit was designed to bring up the **E** data line in these positions only.

A dual NAND gate will produce a 0 only when both inputs are a 1. Studying the address truth table (**table 2**) for what is unique about these two positions of address, it's seen that **B** is, of course, a 1 for both 2 and 6 and adds to **C** for the 6 position. But looking again, **B** is uniquely present on only 2 and 6 if odd numbers involving an **A** can be ignored. This suggests the approach. Invert the **A** and we have a 1 only for even numbers — use it for one gate input with the **B** data line on the other gate input.

Now we have a 0 out of this gate on only the 2 and 6 positions. Since we needed a 1, another section of the 7400 provides the inverter between it and the **E** data line of the S2A.

Taming the 74193. Several conditions must be met for the 74193 to step along through its addresses in sequence, reset at the end of the sequence or sentence, and for its selected data to be presented to the speech module only:

1. When it is valid data
2. When the module is not busy saying the previous word.

Consider first the data provided by the digital display unit (see **table 3**). The count periods are a precise, crystal-controlled, ten milliseconds each followed by 50-millisecond periods of stored display data, alternating about seventeen times a second between the two. We must accept data only during the storage time, or the readings will appear random, having been taken on the "fly" during a count. The waveform that defines these two periods is present

table 1. S2A Speech Synthesizer vocabulary.

data lines up	speech	data lines up	speech
none	OH	C,D	PERCENT
A	ONE	B,C,D	LOW
B	TWO	A,B,C,D	OVER
A,B	THREE	E	ROOT
C	FOUR	A,E	EM
A,C	FIVE	B,E	TIMES
B,C	SIX	A,B,E	POINT
A,B,C	SEVEN	C,E	OVERFLOW
D	EIGHT	A,C,E	MINUS
A,D	NINE	B,C,E	PLUS
B,D	TIMES*MINUS	A,B,C,E	CLEAR
A,B,D	EQUALS	D,E	SWAP

on pin 6 of the 7400 at the right edge of the DD-1 board — the second chip from the back. It is in the logic 1 state during count, 0 during read. We want a signal that is up during the storage period. Since there was an unused section of this 7400, I made an inverter of it by connecting its pin 6 to 9 and 10, bringing out the desired signal to the interface board from pin 8.

Interconnection hints. The grubby details of interconnecting with the DD-1 deserve comment. The picture looking into the top of the unit shows a small terminal board mounted at front right above the 7447s. The circuit was built from a small rectangular piece of circuit board stock, cross-hatched with a hacksaw to provide rows and columns for the data lines. The ventilating holes to the front and right of the power transformer were enlarged to accommodate the twenty-eight interconnecting leads, leaving out the bottom of the accessory unit. I removed the mounting feet and replaced them with spacers, which fasten the DD-1 to the interface unit.

The 100-Hz pushbutton of the DD-1 has a spare spdt section; the three leads added to it and the blanking output previously described are included in the harness.

DD-1 mods

To facilitate connection to the 7447 inputs and provide the blanking output, it's necessary to remove the DD-1 from its case — not a trivial job. Take out the power transformer mounting screws, all screws mounting the circuit board, and remove the input connector, the knob, and bandswitch mounting. Push in the LEDs and the panel should lift out from the back. Make short connections to the added terminal board; the harness may be added after reassembly. A magnifying lamp and small iron and solder are essential. If you decide to drill any holes in the circuit board, for gosh sakes hold the board up to the light first so you won't drill through conductors that may be on the other side.

Modification of the DD-1 is the most miserable part of the project, but finding the patient well after all this surgery should buoy you up for the remaining pitfalls awaiting you.

interfacing the speech module

Now consider interfacing the speech synthesis module. The ROM of this unit is permanently stored with digital data which, when swept out by its internal clock (roughly 12 kHz), produces an audio waveform imitating the twenty-four words of its vocabulary. When properly filtered to reject the dominant clock frequency and amplified, acceptable speech is reproduced by the speaker. The complete program of the S2A is shown in **table 1**. (The unused words may come in handy when debugging the unit.) **A** is the least-significant bit; its pin is 7. **B** appears on 8, **C** on **K**, **D** on **H**, and **E** on pin **F** of its twenty-pin edge connector.

If you're eager to try the speech board before finishing the project, connect all the power leads and indicated ground leads, then tie all data leads either up or down (but don't leave them floating or you'll have problems). Pulsing the start line up with a clip lead, then back to ground, should produce the word you've coded into the speech board by the data line hookup you've chosen.

The **BUSY** line goes to almost −10 when saying words, then goes back to +5 on completion. An emitter follower, with its emitter returned to ground, converts this to the TTL voltage swing required by the interface.

event sequence

Two latches are used in the control loop to guarantee the proper sequence of events. The left-hand latch normally passes on the information that the last word is complete. It then steps the address into the next position. However, when the **D** address line goes high for address **8**, the inverter opens the latch, no more steps are permitted, and the sentence is terminated. If the 100-Hz switch on the DD-1 is not depressed, the lower gate operates at address **6**, cutting the sentence two words short.

table 2. Address truth table.

multiplexer address position	multiplexer address lines A	B	C	D
0	0	0	0	0
1	1	0	0	0
2	0	1	0	0
3	1	1	0	0
4	0	0	1	0
5	1	0	1	0
6	0	1	1	0
7	1	1	1	0
8	0	0	0	1

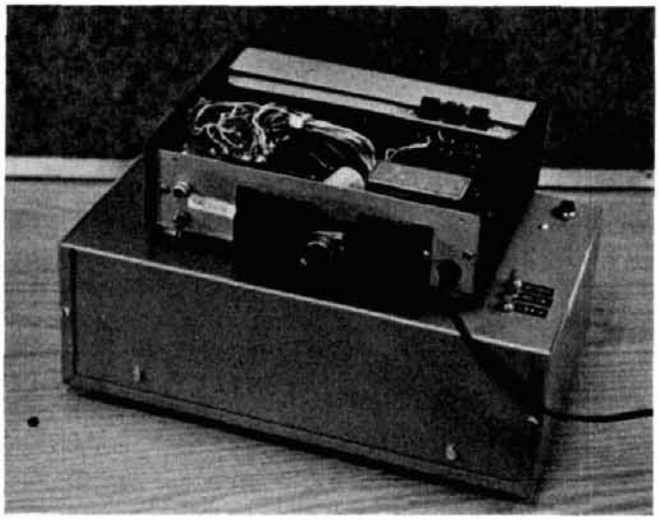
The right-hand latch has a logic 0 entering it at completion of a word, after the address has been stepped, which it passes along and inverts the next time a valid read period occurs from the DD-1. The latch Q output is inverted internally and serves as a start pulse to the speech module, telling it to **GO** while the data is valid.

As I've indicated, the unit normally stops talking at the end of a sentence; this is its initial condition with all address lines high. The pushbutton pulls down the 74193 **LOAD** terminal, resetting it to the 0 address, so it runs through the permitted sequence to address **8** and stops.

Another feature of the 74151 involves the use of its preset capability. If all four preset lines are in the 0 logic state, a down pulse on the **LOAD** terminal will cause it to go to the 0 address. If the **A** preset line is tied high, the reset will go to address **1**. If both **A** and **B** preset lines are tied high, while **C** and **D** are at ground, the unit will reset to address **3**.

This design permitted two useful features to be incorporated. When the DD-1 is presenting 160, 80, or 40 meters, only one digit is needed to enumerate the megahertz. Spectronics connects all four data lines for the first digit high on these bands to cause the digit to blank out. But this presents an **A, B, C, D** to the S2A and it will say OVER in this first-digit location. Since the DD-1 reads only a 1 or 2 when it reads anything in the first-digit position, only the **A** and **B** data lines need be connected to the multiplexers. Connecting the **C** and/or **D** data lines to the **A** preset line will cause the program to skip that digit entirely, which will shorten the beginning of the sentence by one word. The designers of the 74193 must have had this in mind, wow!

Rather than connecting directly to the preset line, however, I ran the **C** and **D** data lines through one section of a dpdt toggle switch so that both the **A**



Back view with the cover of the DD-1 removed. The harness and circuit board are visible at the left. The harness enters the talking readout unit through an enlarged hole adjacent to the power transformer in the DD-1.

its anode on the 74193 LOAD terminal and cathode on the sentence-terminator clock input of the left latch permitted the unit to AUTO-RUN, repeating however short or long a sentence were chosen, ad nauseam. It apparently provides a sufficient down pulse to the LOAD terminal to reset at the completion of each sentence. Really slick for an afterthought!

The interface unit was built on a Radio Shack project board, about 102 mm (4 inches) square with a 44-pin edge connector printed on one edge. There are three rows where up to nine device sockets can be mounted as well as printed tabs permitting two or three connections to each chip terminal. Keeping the data lines in groups of four, along side of the connector, makes tracing easier; the extra one or two can be at the end that didn't quite make it.

table 3. Relationship between the address, displayed frequency, and output of speech module.

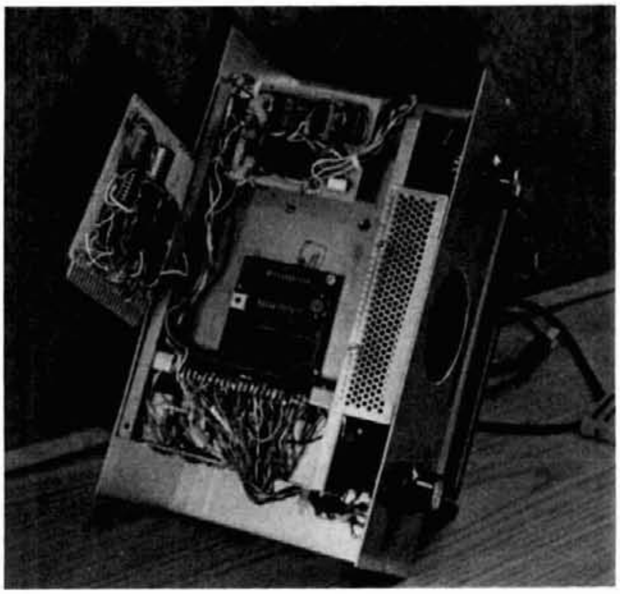
DD-1 display	1	4	•	2	5	0	•	3
speech	ONE	FOUR	POINT	TWO	FIVE	OH	POINT	THREE
address position	0	1	2	3	4	5	6	7

and **B** preset lines can be tied high in the other switch position (see **fig. 1**). The other switch position will delete the ONE FOUR POINT from the start of the 20-meter reading when desired; call it megahertz Delete, I guess. Shortening the sentence to just the three-kilohertz numerals takes only 3/8 as long to say, permitting more rapid checking while tuning around the band with the transceiver. This can, of course, be switched in and out.

While playing around with the unit at this stage of development, I noticed that connecting a diode with

Make a sketch and assign each 74151 a data-line letter: **A**, **B**, **C**, or **D**. One of each four data lines for a given digit will go to the same number pin of each 74151. The first digit lines go to pin 4, second to pin 3, third to pin 1, fourth to pin 15, fifth to pin 14, and sixth to pin 12. Pins 2 and 13 of the **A** and **B** multiplexers go to +5 volts; the same pins of the **C** and **D** multiplexers go to ground as a part of the fixed decimal **POINT** at these address locations.

Wiring can be completed, on the device-side of the board, I recommend about no. 24 (0.5-mm) stranded



Internal view of the talking readout chassis. The speech synthesis module is in the center of the chassis, directly behind the speaker. The interface board has temporarily been removed from its connector. At the top of the chassis is the clock board which has been added to incorporate a talking clock feature. This unit will be described in a subsequent issue of *ham radio*.

wire as about the largest practical size to work with. The picture tells the story; it's a real birds nest! Use the bus down the center row of devices for ground; the other two for +5 volts. The only reason for signals getting where they shouldn't is solder bridges, and if they don't get there at all it's poor connections. After installing all the devices, a couple of point-to-point checkouts with the ohmmeter are worthwhile. It boggles my mind to imagine this as an etched board — perhaps someone will accept the challenge!

power supply

The power supply (**fig. 2**) is conventional and requires little comment. The +5 volt supply uses a 6.3-volt, 1-amp transformer. I slipped in a half dozen extra turns over the existing secondary winding without tearing down the transformer so that its output is close to 7 volts, providing adequate voltage for good regulation. The total load at 5 volts is about 300 mA. A small 300-mA, 12-volt transformer from Radio Shack is adequate for the light load of the negative supply.

The case of the 7805 may be bolted to the chassis; the 7815 can be insulated with a mica or fishpaper shim. To realize a −10 volts, the positive terminal is connected to the +5 volt output; the negative common will then be 10 volts below ground. The full 15 volts of the negative supply is used by the LM380 audio amplifier, so the speaker should be returned to the −10 volt bus.

some afterthoughts

There are other ways of building this unit. Perhaps mounting the four multiplexers on a small board internal to the DD-1 would make things easier and minimize the interconnection problem. This approach shoulg permit a unit no longer than the DD-1 for the balance of the circuit. I hope this is only the first article on the subject. Good luck on your talking second operator!

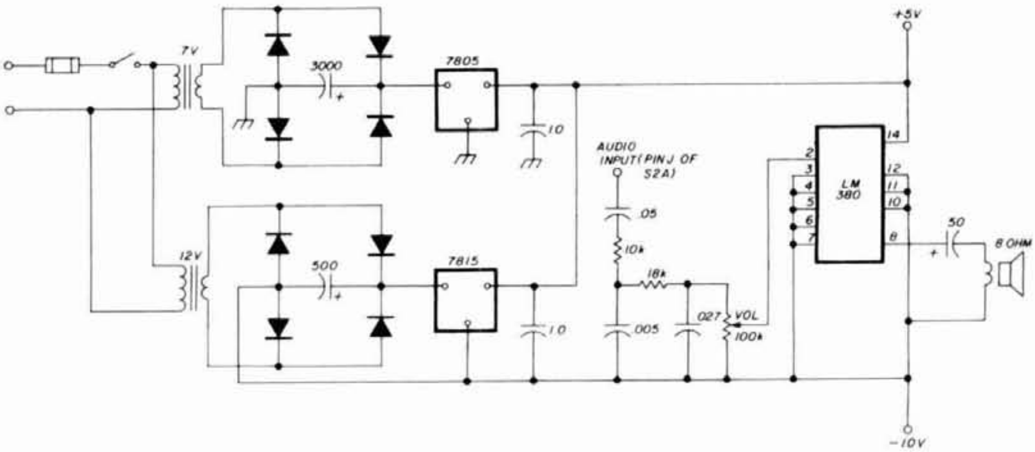


fig. 2. Power-supply schematic. Circuit is conventional. Parts are available from Radio Shack and other popular suppliers.

Oh yes — should you desire to interface a visual readout with only the seven-segment connections to the LEDs available for data, National Semiconductor makes the 86L25, which will convert seven-segment data back to the BCD mode. One would be required for each digit, of course.

ham radio

More Details? CHECK—OFF Page 110

an introduction to packet radio

An interesting idea from our Canadian colleagues: computer time sharing over vhf links on the vhf bands

If ever two hobbies were made for each other, they must be microcomputers and Amateur Radio. More and more hams seem to have taken the plunge and are starting up small computer systems. A logical outcome of this marriage of activities is that sooner or later the ham will wonder if he can't use the radio to send his computer information to his friends on the air. It can be done, and this article describes one of the best ways of doing it.

time sharing

Most computer users have been exposed to the time-shared computer. The computer is very fast, and the users require and generate information very slowly. If each user is connected to the computer by a separate line (**fig. 1**) or a radio link (**fig. 2**), then all the computer has to do is to check each line periodically and divide the processing time among the users. Typically it might take 20 seconds to type in a line on your terminal and only a few milliseconds at most for the computer to process that data line. Time sharing permits each user to think he has the whole computer to himself. It also permits the computer to act as an intelligent clearing-house for programs in which the users interact with each other.

The next step is to look at the lines, or radio links, themselves. The line that took 20 seconds to type might contain 64 eight-bit ASCII characters, or 512 bits of information — an average rate of 512/20, or about 26 bits per second.

It's easy to send 2400 bits per second through a normal voice channel and it's theoretically possible to use much higher rates. So, if we were to store each line locally then send it in a short burst, it could be sent in about 0.2 second! More detailed estimates and calculations show that several hundred users could be accommodated on one voice channel.

Also, because only one frequency (or two for full duplex) is used, newcomers can join the system easily by getting onto that frequency without having to wait for new channels to be assigned at the computer. Now we have a time-shared radio link working with a time-shared computer (**fig. 3**). As we'll see, the time-shared radio link is useful by itself, without the central computer. Its major advantage is a huge saving of spectrum space (not to mention knowing what frequency to look on to find your friends).

packets

Each of the short bursts mentioned above is called a *packet* and most contain, in addition to the data, the identification of both sending and receiving stations and some form of error checking, so that the recipient will know if the information is correct. If it is correct, the recipient sends an acknowledgment (ACK) to the sending station (a fully automatic process). Full details of the packet format appear below.

Transmission of radio packets by Amateurs is now legal on several vhf and uhf bands in Canada. U.S.

By Ian Hodgson, VE2BEN, 296 Malcolm Circle, Dorval, Quebec, Canada H9S 1T7

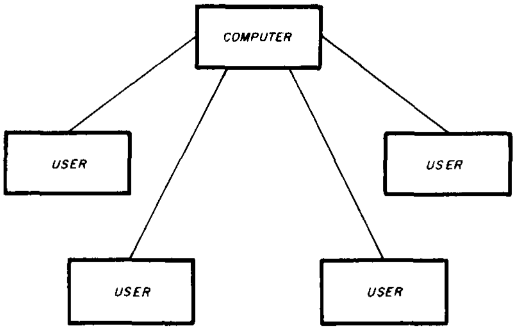


fig. 1. A wired time-sharing system requires separate connections for each user. It permits communications between users as well as computing functions.

Amateurs may want to apply pressure to be permitted to send packets also.

networks

Although two stations can send packets back and forth, the method doesn't really come into its own until many users share a packet repeater (called a *node*). Many nodes can be linked (possibly by a geostationary satellite — this system may actually be working in a couple of years) to form a network (**fig. 4**). Each packet contains its own address information, so the network can automatically forward the information to its destination.

The first radio packet network in use was set up by the University of Hawaii and is called the Aloha net. It links terminals on several islands to a central computer on about 400 MHz.

the node

The repeaters in a network may take several forms. They may be simple rf repeaters used to extend range, as we now use on two meters. This scheme, however, doesn't fully exploit the advantages of packets.

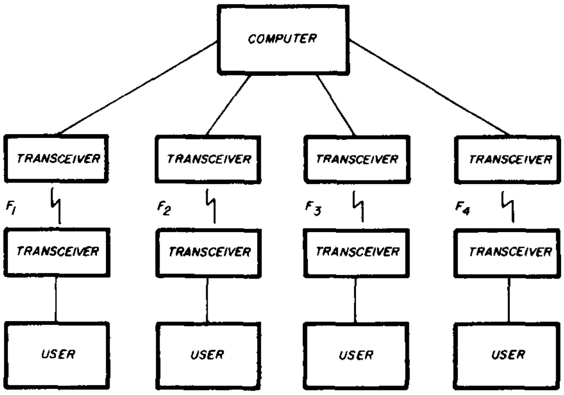


fig. 2. Radio equivalent of a wired system. A different frequency and a separate transceiver at the computer are required for each user.

A better repeater is the *store and forward repeater*, which receives packets from one or more users, processes them, and forwards them to a central node, which would otherwise be too far away to access. If major cities had intelligent nodes, store and forward repeaters would make them accessible to users in outlying areas.

Major nodes would likely be computer controlled to perform error checking, acknowledging, and routing of packets between users. Ultimately, the node might be connected to a central computer with more power than the individual users could afford to have at home. You could have access to utility programs, a ham community bulletin board, repeater council data, a swap program, or even play multi-user space war instead of rag chewing.

what else can it do?

Any information that can be put into digital form can be sent through packets. This information includes RTTY, computer data, and even voice and TV pictures. Of course, data rates must be much higher for the last two modes. The packet system could be as easy, fast, and reliable for talking to hams a continent away as for talking across town on the repeater.

how about details?

Let's take a closer look at the details of the process. The packet format hasn't yet been finally decided upon, but here's what we've been experimenting with (each byte is eight bits).

bytes	contents
1	packet initiation byte (hex A7)
1	start of header (SOH, hex 01)
6	destination call sign
2	destination node
6	originator's call sign
2	originator's node
1	service message flag byte (set to hex 41, A, for ACK)
1	spare
1	end of header (EOH, hex 04)
2	CRC16 (this is a type of error check) for header
(Everything above is the header)	
1	start of text (STX, hex 02)
64	DATA (could be ASCII, binary or EBCDIC)
1	end of text (ETX, hex 03)
2	CRC16 for data
91	bytes total

Note that all packets must be the same length if the system is to operate efficiently (if you don't understand why, let's just say that it can be shown mathematically, and, even though I can do the figures, I don't really understand either).

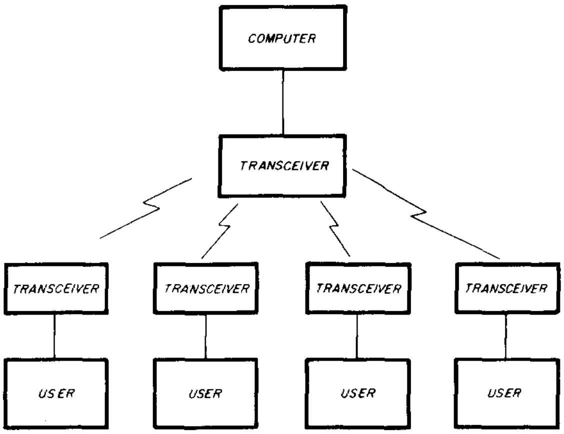


fig. 3. A packet network needs only one frequency and one main transceiver. In addition to computing power, the network offers flexible communications between users, a fact that Amateurs can put to good use.

An ACK is simply the header from the received message with the originate and destination addresses interchanged and with the service byte set to indicate that it is an ACK. No data is included in an ACK.

interference and ACKS

It will undoubtedly come to pass that two users will, sooner or later, send their packets at the same time, thus causing total or partial loss of both. We can deal with this problem by using what's called a CSMA POSACK system. This acronym stands for Carrier Sense Multiple Access Positive Acknowledgment. Here's how it works:

You finish typing your line of data and hit "return," or whatever, to send the packet. Your system checks with your receiver, and if no signal is being received, the system immediately sends the packet. It then waits for a time of one packet length for the ACK. If not received, the system waits an additional, random, number of packet lengths and tries again. The process is repeated three times. If still no ACK is received, the system returns a message to the sender, telling him that the transmission was unsuccessful.

Why do we wait a random delay? If there were no random delay, and two stations sent their initial packets at the same time, then all three tries would collide. This way, they don't.

When receiving packets, your system intercepts all packets regardless of their address. The system checks the destination of every packet, and if it's for you, the error check (CRC 16) is performed. If this is OK, the system immediately sends an ACK. Note that

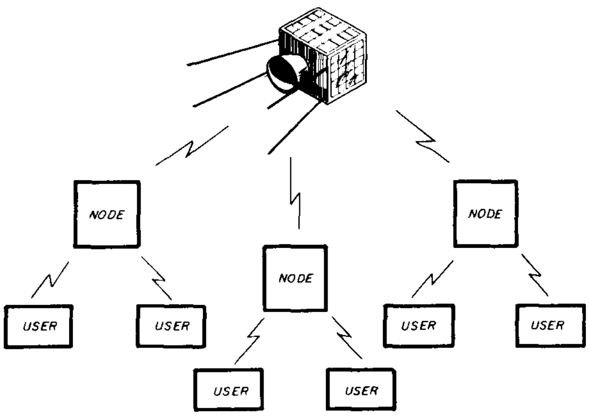


fig. 4. A packet network allows time-shared communications between users and a variety of nodes. Nodes in various cities may be connected together by uhf links or perhaps by satellite. Not as far out as it seems, this scheme may be in operation within two years.

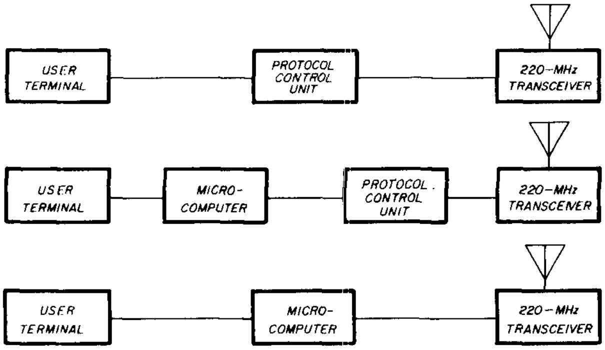


fig. 5. Three possible packet radio setups. The protocol control unit may be used by itself to implement the packet protocol, or it may be used in conjunction with a microcomputer to allow computer information exchange between users. Alternatively, the microcomputer may be programmed to take care of the protocol itself, with some reduction in its availability for other purposes.

ACKs have priority on the system, as all transmitters wait one packet length after each packet sent so the ACKs can get through. Also if a signal is present on the initial receiver check, the packet transmission from your station is delayed long enough so that the other station can be acknowledged.

hardware

Sounds complex? Well, it is. But don't forget how inexpensive complexity is becoming; witness the pocket calculator, which for less than $10 is more complex than everything else in your house put together. Here's what you really need:

a. A vhf (220-MHz in Canada) transceiver (fm will do)

b. A computer, Protocol Control Unit, or both (see **fig. 3**)

A single-board microcomputer can form the Protocol Control Unit (PCU) to implement all the above details (these details are called the packet *protocol*). It would require only about 2k of memory and could probably be built for less than $100. We are working on this too. In due course, these should be available commercially, but for now let's build our own. Full details of the final protocol will be published when available.

what frequencies can I use?

In Canada certain portions of some vhf bands have been set aside for packets:

frequency (MHz)	transmission mode
220.1-220.5	shared packet and other modes
220.5-221.0	shared wideband packet and other modes
221.0-223.0	packet only
223.0-223.5	shared wideband packet and other modes
433.0-434.0	packet only (wideband, 100 kHz, for repeater links)
24,000-24,010	shared packet and other modes

Obviously the bulk of the activity is expected to be on 220 MHz. If you've looked at the low prices on rigs for that band, it seems like an excellent idea.

how it all began

In April, 1978, the Department of Communications (DOC) in Canada announced its intention of changing the regulations to permit only "Packet Radio" (cries of, "What the devil is that?") on 220-225 MHz. Amidst the horrified screams of many Amateurs, a few of us decided to have a closer look. We liked what we saw. With the close cooperation of the DOC, who modified their original proposal accordingly, we started experimenting with packets. The group doing most of the work is based in Montreal.

One more thing to note. The Amateur Radio community is the group that will develop this system on a low-cost, widely distributed basis. Our work will no doubt be closely watched by commercial interests, so let's earn our privileges as hams and contribute once more to the advancement of the state of the art.

acknowledgments

I'd like to thank Bob Rouleau, VE2PY, for enthusiastically supporting the packet concept and keeping the rest of us going; Paul Laflamme for his work on the protocol and programming and for helping to prepare this article. Special thanks to Dr. John de Mercado, Director General of the Telecommunications Regulations Branch of the DOC, for giving us the initial kick by changing the regulations and for his continued support and close cooperation. I wish *all* government departments worked this way!

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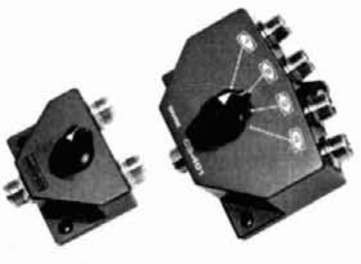

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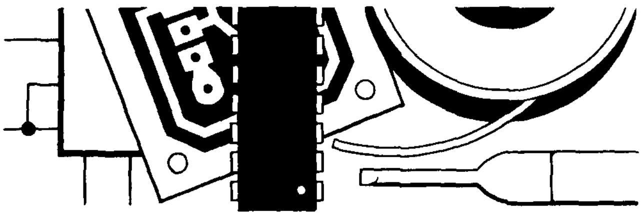
the weekender



biquad bandpass filter for CW use

The biquad bandpass filter has many advantages when used as a CW filter. First, the biquad can be adjusted to control both center frequency and *Q*. Second, the filter is very stable and will not break out into ear-shattering oscillations. Third, the filter can be built by hams like you and me. That is to say, the components are available at Radio Shack, the tolerance of none of the components is critical, and no special test gear is needed.

circuit description

Fig. 1A shows the traditional circuit, which requires the plus and minus supplies. The advantage of this configuration over the single-voltage circuit is the capability to couple directly to the input and output of the filter. In addition, the op amp has better gain characteristics with the higher supply voltage.

The single supply circuit is very similar (see **fig. 1B**). Most references recommend providing a low-impedance dc return for the input circuit. For this reason, I avoided for a long time even trying the single-voltage supply. But my experiments with the μA741 showed that the circuit does a remarkably good job. By providing half the supply voltage to the positive input of the op amp, all of the outputs are operating above ground. This is the obvious reason for needing the capacitive coupling on the output and input. The μA741s will drive headphones directly, but for general use, an amplifier speaker is recommended.

By James M. Rohler, NØDE, 1967 Bristol Drive, Bettendorf, Iowa 52722

After a short period of experimenting, you will find the adjustable features useful. The *Q* is especially interesting, since it can be adjusted too high, causing ringing. Under some conditions, the narrower band-

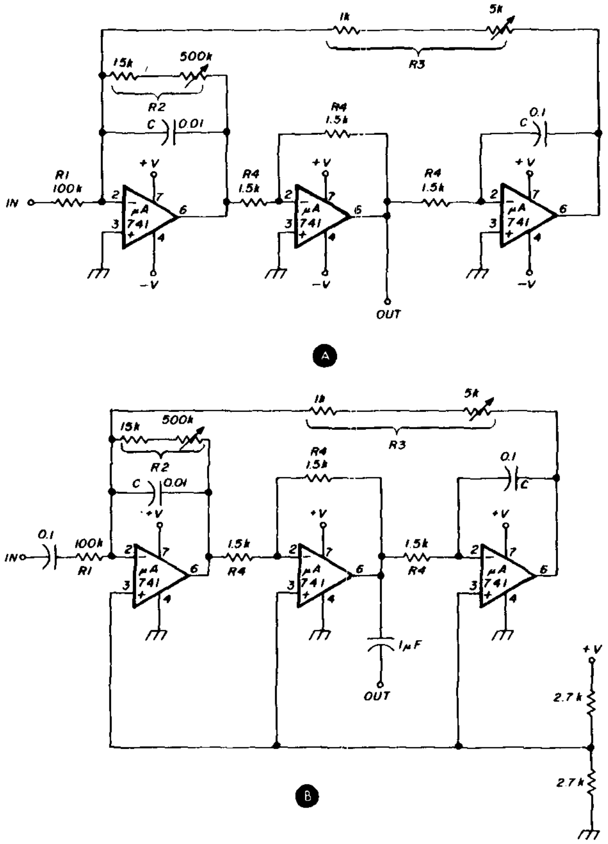


fig. 1. Schematic diagrams of the biquad active filter for dual supply (A) and single supply (B).

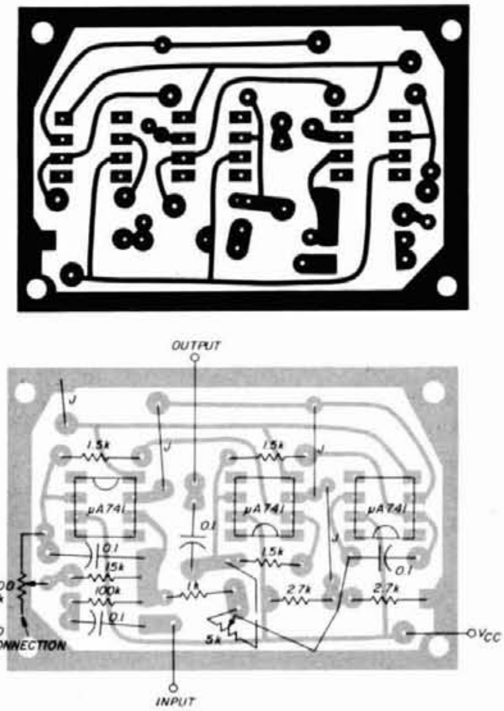


fig. 2. Foil pattern for the active bandpass filter (above). The parts placement diagram is shown (below).

width and ringing is a fair tradeoff for eliminating the QRM.

components and construction

Fig. 2 shows the foil pattern and parts placement diagram. If potentiometers are not desired, fixed resistor values can be determined from the following equations:

$$f_o = \frac{1}{2\pi C\sqrt{R3R4}} \quad (1)$$

$$B = \frac{1}{2\pi CR2} \quad (2)$$

$$G = R2/R1 \quad (3)$$

where f_o is the center frequency in Hertz,
B is the bandwidth in Hertz,
G is the gain of the circuit.
The Q is the reciprocal of the bandwidth times the center frequency.

For single-supply operation, jumpers are used to connect the junction of the bias resistors to pin 3 of each IC. In addition, pin 4 of each IC is connected to ground. To use a dual supply, the pin 3s are connected to ground, and pin 4s to the minus supply. Also, in this case, the input and output capacitors and bias resistors can be eliminated.

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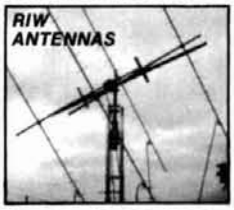

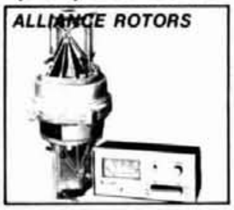

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A220-11 11 ele 220MHz beam . . . . **34.95 c**
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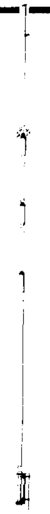

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gallon-size dummy load

Homebrew dummy load with low SWR that includes rf-voltage monitoring provisions

In a previous article,[1] *I described a home-built* dummy load based on intelligent overload of a bank of carbon resistors good for 5-second tuneup of a 1-kW transmitter (50 per cent loaded at tuneup). The dummy load served me well for many years and, in fact, is still in use. But an incident occurred recently during a TVI test of a new linear that emphasized that a rating of 500 watts for 5 seconds doesn't mean a full kilowatt continuously! A new set of resistors got the load back to normal, but the incident caused some time out to build the dummy load described here, which is capable of running tests at the 2-kW input level.

review of rating techniques

The following applies to Allen-Bradley 2-watt resistors. Other sizes, or resistors of other manufacturers, can be used, but the rating coefficients must be obtained from the manufacturer.

A 2-watt, A-B resistor of the 20 per cent series is rated for a continuous load of 2 watts for 100,000 hours when the resistor is mounted with 25.4-mm (1-inch) leads and has a body temperature of 100C (212F). This occurs when the ambient temperature is 50C (122F). The life rating increases by a factor of ten for a 50C (122F) reduction in temperature. The allowable load increases by 40 per cent for a 10:1 reduction in life. For short-term loads the resistors are rated at 44 watt-seconds for the same mounting and ambient temperature, *i.e.*, 44 watts for 1 second, and so on.

The required life for a test dummy load is very short; a few hundred hours in many years. Also, good cooling can be provided by mounting resistors with metal fins touching the resistor body and immersing the resistor body and fins in oil. Even after considerable testing, the resistor body temperature need not exceed 50C (122F).

The shorter life requirement allows the power input to be increased by a factor of 1.4^4 and the lower temperature by a factor of 1.4. The rating now becomes 10.6 watts continuous, or 47 watts for 5 seconds. It's somewhat over 28.5 watts for 10 seconds and well over 20 watts for 20 seconds. (Ambient temperature isn't a major factor for very short loads, but resistance change is.)

At high power there's another factor to consider. The maximum voltage rating of a 2-watt resistor is 500 volts, which will affect the mounting method.

choice of design values

Since some work is done at full 2-kW input, a dummy load that could accept this power for short test periods seemed desirable. A reasonable assumption for maximum linear efficiency is 60 per cent, so the desired rating was 1200 watts for 5 seconds, which would allow 300 watts continuously.

The voltage rating must be reduced by the peak-to-average ratio of the applied signal. Also, for high frequency power use some dielectric heating of the resistor body will occur. With some allowance for this, the rf voltage across a resistor should not exceed 250 volts rms.

At this voltage the minimum resistance for a peak dissipation of 47 watts is about 1325 ohms. A total of 26 resistors in parallel would be needed for a 50-ohm load. However, since this is not a standard value, some adjustment is necessary.

The local surplus emporium had no 2-watt resistors close to the 1200-ohm value, but there was a large bin of 470-ohm units at a very good price. A quick calculation showed that eighteen of these in parallel would give a resistance of 25 ohms, with two such banks in series giving the desired 50 ohms. Alternatively, two banks of fifteen resistors each

By R. P. Haviland, W4MB, 2100 South Nova Road, Box 45, Daytona Beach, Florida 32019

would give 60 ohms, nearly the mean between 50 and 75 ohms. The last combination was chosen for construction.

dummy-load circuit

While the dummy load is usually used with a wattmeter, an independent power check is sometimes useful. A built-in rf voltmeter can give this measurement. Because of the high voltage present, a voltage divider must be included. The circuit is shown in **fig. 1**.

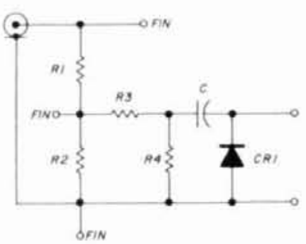
With the resistors in hand, the mounting and oil immersion problems must be worked out. For very short periods of operation, a 0.946-liter (1-quart) can would be suitable. However, the temperature rise is rapid, and more oil is desirable, so 3.8 liters (1 gallon) allows reasonable test periods.

construction

It's easier to show the construction than to describe it. **Fig. 2** shows the load assembly. Three fins provide the connections for the two banks. One outer fin soldered to the can case provides most of the assembly support and the connection to the outer coax lead. The other outer fin is drilled to fit over the center conductor of the coax receptable, providing the remainder of the support. (Note that the three fins should be the same size to keep the voltage equal across the banks.) The voltage-monitoring components are mounted on a terminal strip with tip jacks used for the output.

If possible, use hermetically sealed connectors. Transformer oil, the preferred cooling medium, tends to migrate, and unsealed connections will always be oily. Even light mineral oil will migrate. Hydraulic brake fluid, the other possible coolant, also migrates. If you must use standard connectors, be generous with silicone rubber sealant.

Performance of the unit is good. The SWR is about 1.2:1, with a 50-ohm wattmeter, to above 30 MHz. However, at higher frequencies the uncompensated reactances affect performance. The apparent resistance on 144 MHz is appreciably reduced, and the SWR is undesirably high. While the unit is usable at vhf, it's basically a high-frequency design.



R1,R2 EACH 470 OHM 2 WATT, ALLEN BRADLEY RESISTORS IN PARALLEL
R3,R4 1k 1 WATT
C 0.001 μF CERAMIC DISC
CR1 GERMANIUM DIODE

fig. 1. Dummy-load schematic, including voltage-monitoring provisions.

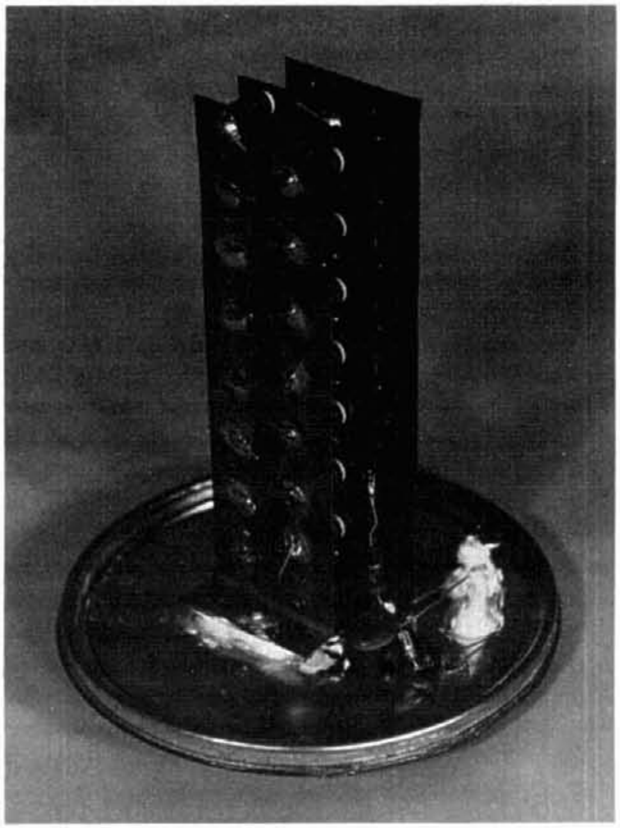


fig. 2. Construction of the gallon-size dummy load. Fins are made from copper flashing. The 2-watt resistor leads project through holes in the copper, are bent over, and then are soldered. The fins should touch the resistor body. Resistor spacing should be two body diameters or more. The fins should clear the can sides by 12.5 mm (½ inch) or more.

variations

Many variations of the basic design are possible. Sixty resistors would handle a kilowatt transmitter for long periods. Exact 50- or 75-ohm loads can be worked out for SWRs around 1.05:1. Special resistances are possible, such as 8 ohms for audio amplifier tests. Stray reactance can be tuned out for vhf operation.

The unit shown has been operated many times with a linear amplifier running at 1-kW input for periods of 10 minutes or so. No measurable change in resistance values was found. The unit has paid its way: the linear is now free of TVI thanks to the testing time possible.

reference

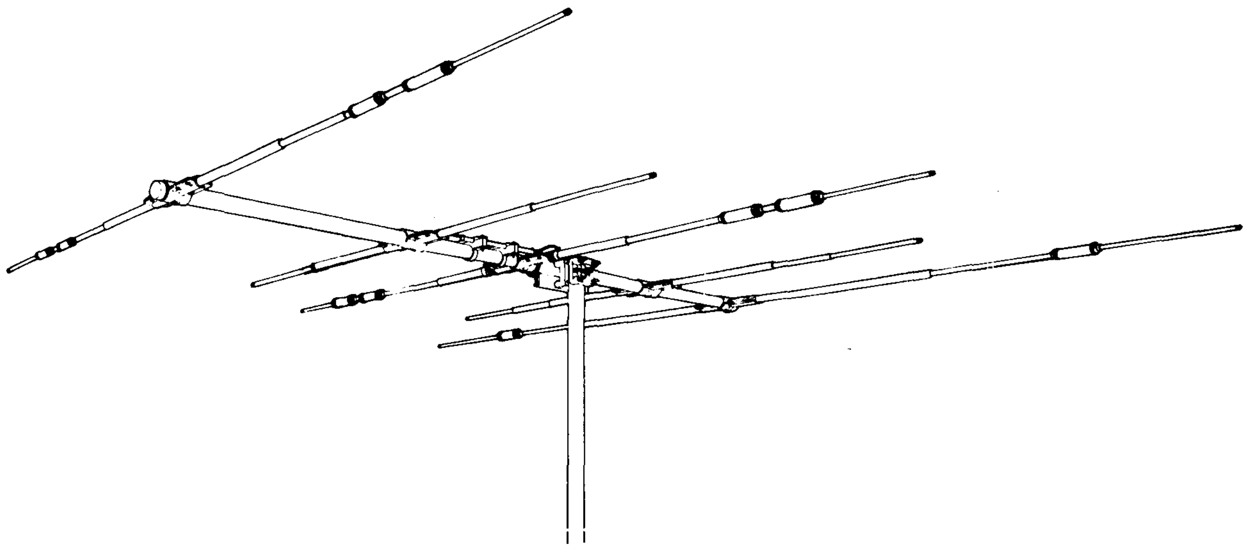
1. Robert P. Haviland, W4MB, "Superfluous Signals," *ham radio*, March, 1976, pages 40-43.

ham radio

More Details? CHECK—OFF Page 110

TH5DX

10-15-20 METERS



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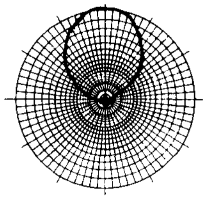
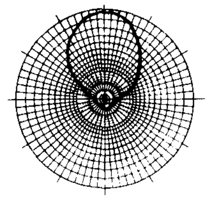
Boom length .... **18 feet**
Longest Element .... **31 feet**
Turning Radius .... **18 feet**
Surface Area .... **6.4 sq. feet**
Wind load .... **164 lbs.**
Weight .... **50 lbs.**

VSWR at resonance ..... **less than 1.5:1**
Power Input .... **Maximum Legal**
Input Impedance .... **50 ohms**
-3dB Beamwidth .... **66° average**
Lightning Protection .... **DC ground**
Forward Gain .... **8.5 dB**
Front-to-Back Ratio .... **25 dB**

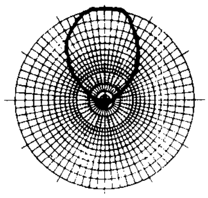
20 Meter



15 Meter



10 Meter



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More Details? CHECK—OFF Page 110

DIGITAL TECHNIQUES

multivibrators and analog input interfacing

Previous parts of this series have concentrated on only one multivibrator, the bistable, or flip-flop. This part will examine the monostable multivibrator, or one-shot, and look into ways of converting analog signals to digital levels.

A one-shot will create a single pulse of controllable width in response to a trigger. The trigger can be either a positive-going or negative-going state transition. There are two basic types: conventional and retriggerable. The latter will automatically reset and start the pulse again if a trigger arrives before the pulse is complete. Most of the former will inhibit any retriggering if the pulse has already started.

A one-shot symbol is shown in **fig. 1**. It has Q outputs and direct set and clear inputs as in a flip-flop.

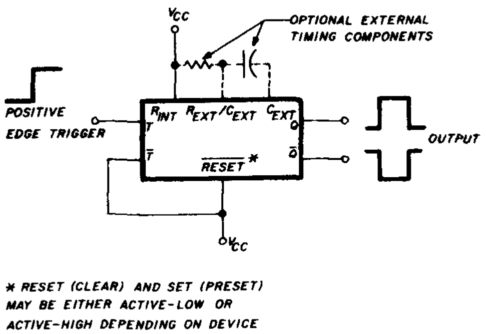

fig. 1. One-shot multivibrator symbology with optional external components.

Some devices allow a choice of positive or negative triggering. As in the flip-flop, unused control and input lines must be tied high or low depending on function.

Output pulse width depends on an RC time constant, either internal or external. The device is partly analog in operation, and a full description is found in texts.[1] Each one-shot IC is specified in width from a simple formula or a graph showing width versus RC combinations. Most all one-shots have internal resis-

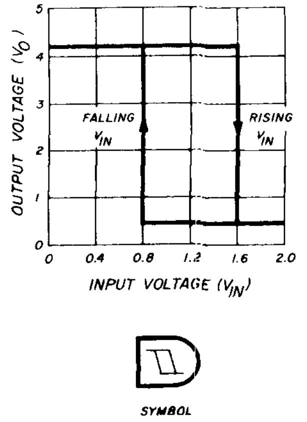

fig. 2. Schmitt trigger gate hysteresis curve and symbol.

tors and capacitors that can be changed by external components. Most confusion in RC selection comes about by the three pins marked R_{int}, C_{ext}, and R_{ext}/C_{ext}. Connection rules are:

Internal C only:	C_{ext} pin left open
External C:	Capacitor between C_{ext} and R_{ext}/C_{ext} pins
Internal R only:	Connect R_{int} pin to V_{cc}
External R:	R_{int} pins left open, resistor between V_{cc} and R_{ext}/C_{ext} pin

Internal capacitance and resistance are specified for each device. External components should be mounted as close as possible to the package; R and C pins are very susceptible to noise pickup, even with TTL.

External resistance may be variable for adjustment. The trimming potentiometer must not be wirewound for short pulses; winding inductance will change the time constant.

astable multivibrators

These are free-running oscillators with digital level outputs. Their use is generally restricted to digital VCOs (voltage-controlled oscillators). Most are of the emitter-coupled variety for stability.* Details can be found in application notes or reference 1.

Most of today's timing is obtained from quartz crystal or LC oscillators, with or without dividers for lower frequencies. Such oscillators don't have the fast rise and fall times required for digital circuits. TTL

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*Not to be confused with ECL, or emitter-coupled-logic.

devices will not work properly with transition times of about 5 microseconds or more; internal circuitry will actually oscillate while a transition is made from logic 0 maximum to logic 1 minimum. CMOS is a bit more tolerant. There are two solutions.

the Schmitt trigger

A Schmitt trigger is a high-gain amplifier with feedback to give hysteresis. Hysteresis is a nonlinearity between input and output and, in digital form, helps to discriminate signal and noise when feeding digital inputs.

Some gates and inverters incorporate Schmitt trigger circuits at each input. So do a few one-shots. Typical input versus output is shown in **fig. 2** for a TTL gate. The hysteresis symbol is used in gates and inverters having Schmitt inputs.

A rising input voltage will cause the output to "snap" from high to low when it crosses the 1.6-volt threshold. Internal feedback allows a very slow threshold crossing to change the output very rapidly. Schmitt inverters and gates are very good for interfacing an analog signal within input maximum voltage swing to conventional digital circuitry.

Other uses are given in **fig. 3**. Circuits of **figs. 3A** and **3B** are useful for automatically resetting a digital circuit during power-on. **Fig. 3C** may be used to debounce a single-throw switch, but with some caution. Values are for TTL, and closure time constant is much shorter than opening time constant. It will work well with CMOS, where resistor values can be equal and much higher.

comparators

Comparators are high-gain, wideband operational amplifiers with output clamps to restrict output voltage swings to digital levels. A typical application is shown in **fig. 4**.

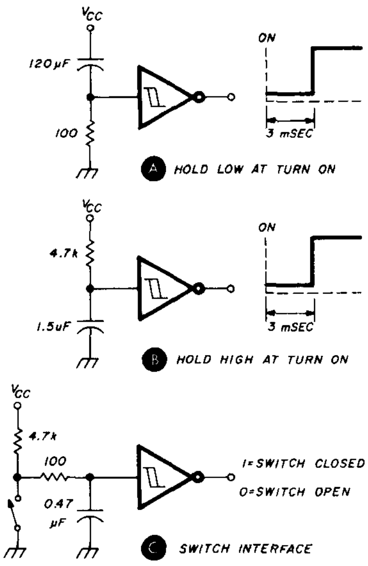

fig. 3. Schmitt trigger applications with inverters.

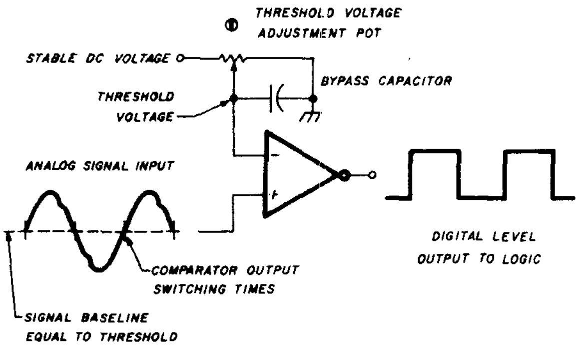

fig. 4. Typical comparator circuit.

Comparators usually need two more supply voltages. This requirement is offset by the ability to threshold signals with a dc baseline way off digital voltage limits. Input voltage thresholds and baseline shifting are the same as for lower-frequency operational amplifiers.

The main characteristic in choosing comparators is slew rate, or response time. Slew rate is rate of change of output voltage per unit time compared with input voltage. Slew rate should be as high as possible. The newer rating is response time and is found on data sheets as time-related graphs of output voltage change for different overdrives.

Overdrive occurs when the input voltage exceeds threshold voltage. Most high-speed comparators have nearly the same output voltage transition with any overdrive; the major change is a slight delay with small overdrives.

Packages such as the Motorola MC3430 have four separate comparators. The MC3430 needs only two voltages, +5 volts and −5 volts, and also has a common strobe control line. All four outputs are TTL-compatible and are three-state. When the strobe is high all outputs are in a high-impedance state, which allows wiring several outputs in parallel. Feedback may also be added for hysteresis.[2]

The Harris HA-4905 is also a quad comparator with an interesting supply connection. Two package pins are used for the comparator section and two more are used for the output circuits. One or two supplies can be used for the comparator with a differential of 5-15 volts. The output-circuit supply may be set to equal the digital-circuit supply for signal compatibility.

references

1. Jacob Millman and Herbert Taub, "Pulse, Digital, and Switching Waveforms," Chapter 11, McGraw-Hill Book Company, 1965.
2. *Motorola Semiconductor Data Library*, Volume 6, Series B, Motorola Semiconductor Products, Inc., 1976, page 6-23.

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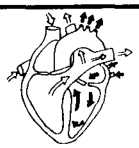

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FOX-TANGO continues to expand its quality-line of 8-pole filters, made up entirely of specially-treated, high Q crystals. Custom-made for each set with ultimate rejection of over 80 dB and superior shape factors, these filters are unavailable from other sources. Only **FOX-TANGO** gives owners of the most popular brands of ham gear a choice of drop-in filters closer to their needs than the limited number of 6-pole (or less) units offered as extra-cost options by most manufacturers. In addition, for the Yaesu and Kenwood lines we offer specially designed **Diode Switching Boards** which permit inboard mounting of up to two filters more than those for which the manufacturer provides room. Thus, either now or in the future, a set could be provided with a sharp (250 Hz) and/or standard (500 Hz) CW filter; and/or a narrow (1800 Hz) SSB in addition to the regularly supplied SSB filter — all switch selectable. This provides superior variable bandwidth selection without the need for buying an expensive new model to provide it. *In these inflationary times, it is sound economics to up-grade your present set.* **FOX-TANGO** makes it easy — and relatively inexpensive. Order with confidence: satisfaction is guaranteed on a money-back basis. Incidentally even our current line of filters will fit many other brands or home-brew rigs. Check the specifications or write us about custom-built units for most rigs.

FOX-TANGO CRYSTAL FILTER SPECIFICATIONS

		FILTER NO. YF-	USED FOR	CENTER FREQUENCY kHz	BANDWIDTH — Hz −6 dB	BANDWIDTH — Hz −60 dB	INSERTION LOSS dB	TERMINAL Z (IN-OUT) Ω/pF	CASE SIZE SEE BELOW	SEE NOTE
YAESU	FT-101 FR-101	31H250	CW-N	3179.3	250 ± 50	<750	<9	500Ω	A	1
		31H500	CW	3179.3	500 ± 50	<1200	<7	500Ω	A	2
		31H1.8	SSB-N	3180	1800 ± 100	<3100	<6	500Ω	B	3
		31H2.4	SSB-F	3180	2400 ± 100	<4200	<6	500Ω	B	4, 6
		31H6.0	AM	3180	6000 ± 500	<11K	<6	500Ω	A	5
	FT-301 FT-7	91H250	CW-N	8999.3	250 ± 50	<750	<10	500Ω	C	1
		91H500	CW	8999.3	500 ± 50	<1400	<8	500Ω	C	2
		90H1.8	SSB-N	9000	1800 ± 100	<3100	<6	500Ω	C	3
		90H2.4	SSB-F	9000	2400 ± 100	<4200	<6	500Ω	C	6
	FT-401	31H250	CW-N	3179.3	250 ± 50	<750	<9	800/5KΩ	J	1, 15
		31H1.8	SSB-N	3180	1800 ± 100	<3100	<3	800/5KΩ	J	3, 15
		31H2.1	SSB	3180	2100 ± 100	<3100	<4	800/5KΩ	J	9, 15
	FT-901 FT-101Z	89H250	CW-N	8988.3	250 ± 50	<750	<10	500Ω	C	1
		89H500	CW	8988.3	500 ± 50	<1400	<8	500Ω	C	2
KENWOOD	TS-520 R-599	33H250	CW-N	3395	250 ± 50	<750	<9	4.7K/33pF	B	1
		33H400	CW	3395	400 ± 50	<1200	<8	4.7K/33pF	B	7
		33H1.8	SSB-N	3395	1800 ± 100	<3100	<6	4.7K/33pF	B	3
	TS-820	88H250	CW-N	8830.7	250 ± 50	<750	<10	470/5pF	D	1
		88H400	CW	8830.7	400 ± 50	<1400	<9	470/5pF	D	7
		88H1.8	SSB-N	8830	1800 ± 100	<3100	<6	470/5pF	D	3
HEATH	All Except SB/HW 104	33H250	CW-N	3395.4	250 ± 50	<750	<9	2KΩ	E	1
		33H400	CW	3395.4	400 ± 50	<1200	<8	2KΩ	E	8
		33H2.1	SSB	3395	2100 ± 100	<3100	<6	2KΩ	E	9
DRAKE	R-4C	56H8.0	CW/SSB	5645 ± 150	8000 ± 500	<13K	<3	500Ω	F	10, 11
		56H800	CW	5645.5 ± 50	800 ± 100	<1800	<5	500Ω	C	10, 12
		56H125	CW-N	5695 ± 50	125 ± 50	<350	<13	50Ω	G	10
COLLINS	75S-3B/C	455H250	CW-N	455 ± 50	250 ± 50	<750	<10	5KΩ	H	1, 14
$	All filters, except Collins and Drake, are $55. Collins filters are $150 each; for Drake prices see adjacent column. Prices include airmail postpaid to U.S., Canada and Mexico. For overseas airmail, add $3.									

CASE SIZES
(LWH in mm)
A 45×23×28
B 50×25×25
C 40×20×21
D 50×18×18
E 57×24×25
F 57×24×18
G 57×24×21
H 78×16×22
J 30×30×75

NOTES
1. Sharp CW Filter for DX and Contest work
2. Use instead of optional 600 Hz (6 pole) unit
3. For narrow SSB to reduce QRM
4. Use instead of XF-30A (6 pole) in early units
5. Superior to XF-30B (6 pole) AM unit
6. Used in G3LLL RF Speech Processors
7. Use instead of optional 500 Hz (6 pole) unit
8. Use instead of optional 400 Hz (4 pole) unit
9. Superior replacement for standard SSB unit
10. For detailed description and prices see →
11. Also known as GUF-1 — $65 ppd.
12. Also known as GUF-2S — $90 ppd.
13. Sharpest CW filter for contest work, $100 ppd.
14. Introductory price $150 ppd. Limited time only.
15. Applies to series 400, 401, 560, 570.

See you at the Atlanta HamFestival, June 16, 17!

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Specify desired unit completely when ordering. Include make and model of set, filter number desired, and center frequency.

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YF56H8.0 (GUF-1). Broad first IF Filter. Superior 8-pole type (Shape factor 1.5, ultimate rejection >100dB) *replaces original unit (Shape factor 4.7, ultimate rejection 65dB). Used in CW/SSB modes.* Drop-in replacement of original unit — only two soldered connections. Recommended for use in conjunction with GUF-2. With complete instructions. **GUF-1 $65** ppd. Add $3 for overseas airmail.

YF56H800 (GUF-2). Narrow first IF Filter. (Shape factor as low as 2.0, ultimate rejection >100dB.) *Both GUF-1 and GUF-2 improve dynamic range, reduce intermod due to overloading of second mixer,* reduce QRM, etc. GUF-2 includes a relay kit and all parts needed to provide automatic selection of GUF-1 *or GUF-2 depending upon the position of the* Mode Switch, thus retaining phone capabilities lost if GUF-2 (CW) is used alone. Complete step-by-step instructions simplify installation. **GUF-2**, including relays and parts: **$90** ppd. Add $3 for overseas airmail.

YF56H125. Sharp 125Hz 8 pole, second IF Filter. (Shape factor as low as 2.0, ultimate rejection >100dB.) *Sharpest CW filter available.* See Specs in Table. If you can't get it with this filter, you can't get it — period! Puts selectivity where it counts: in the AGC loop. Eliminates loss of RX gain due to QRM outside of passband, commonly noted with audio filters — yet works well with them. Simple installation; plugs into accessory socket on back of the set. **$100** ppd. Add $3 for overseas airmail.

GUD Product Detector Kit. All parts and instructions needed to convert the existing detector to a double-balanced type to provide better isolation between input and output. Overcomes distortion in audio output and AGC attack. Includes compact PCB with active DBM SN16913, relay switch, and accessory components. Detailed instructions, diagrams and photographs simplify installation. **GUD Kit $45** ppd. Add $3 for overseas airmail.

Dealer Inquiries Welcomed
Florida residents add 4% (sales tax)

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More Details? CHECK—OFF Page 110

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- Antenna
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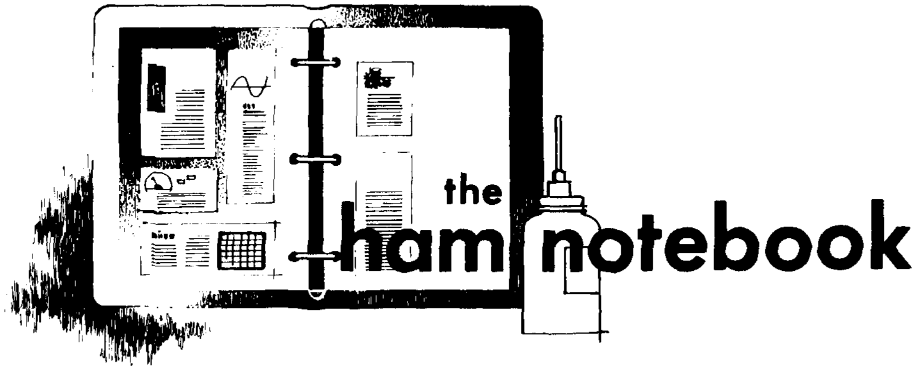

increased break-in delay range for the Heathkit HW-8

The Heathkit HW-8 transceiver break-in delay time constant is such that, even with relatively fast keying and maximum time delay set in, the rig returns to the receive mode after every word. This is fine for rapid break-in, but most CW operation is not rapid break-in, and all that's accomplished is unnecessary relay wear and operator fatigue. The fix is simple: Change C92 from 10 µF to 50 µF, from 10 µF to 50 µF, the transmitter and sidetone monitor oscillator keying develop tails because of the coupling back through the solid-state devices. This side effect may be eliminated in the transmitter keying circuit by lifting the end of R64 nearest the back of the printed circuit board, where wire "W" connects, soldering the anode lead of a 1N4148 silicon diode (Radio Shack 276-1122) in the

In addition, receiver muting is delayed excessively by the increased size of C92. This may be corrected by placing a 4700-ohm resistor in parallel with C43. Remove C43 from the circuit board, solder it across a 4700-ohm resistor, and place the pair where C43 was originally installed. (Refer to **fig. 1.**) All resistors are ¼ watt.

After making the modifications,

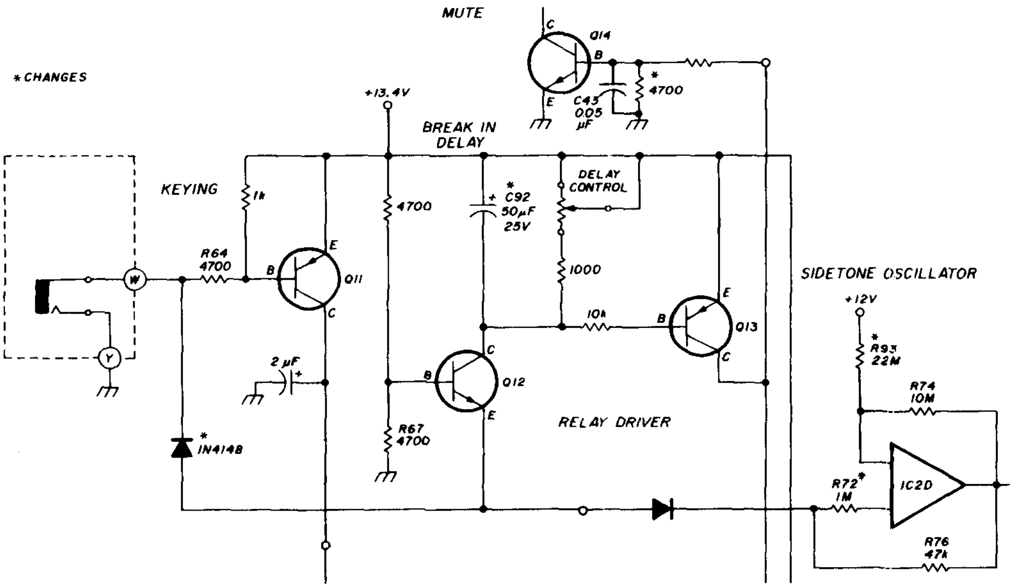

fig. 1. Changes to the Heathkit HW-8 for increased break-in delay range.

25 volts; R72 from 3.3 megohms to 1.0 megohms; R73 from 10 megohms to 22 megohms; parallel C43 with 4700 ohms; and add a diode in series with R64 (see **fig. 1**).

The time constant is controlled by C92. However, when C92 is increased empty hole, and soldering the cathode lead of the diode to the unconnected end of R64. Reconnect wire "W" to the junction of the diode and R64, above the board. In the sidetone oscillator circuit change R72 and R73 as indicated above.

the DELAY CONTROL potentiometer will provide a wide range of time constants. Each operator should adjust the DELAY CONTROL to suit his particular keying speed, with high speed requiring the least delay time.

John Abbott, K6YB

Collins 516F-2 high-voltage regulation

Shortly after acquiring my Collins S-line equipment, I became aware that the CW waveform as viewed on the SB-610 monitor scope left something to be desired. A definite trough appeared in the waveform, which is characteristic of a high-voltage power supply with insufficient regulation (**fig. 2**). The as-built, 10-μF filter is sufficient for SSB operation. However, during key-down CW operation, with the amplifiers running full-bore and exhibiting a widely varying load, there's a need for increased capacitance in the filter section. I found that a minimum of 25-μF of filter capacitance provided the smooth waveform shown in **fig. 3**.

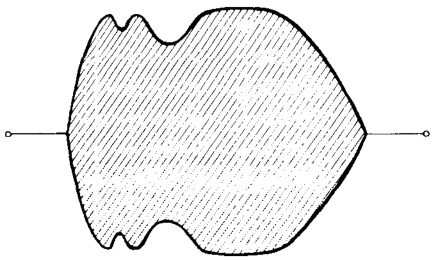

fig. 2. Monitor scope display of Collins S-Line CW waveform showing insufficient high voltage supply regulation.

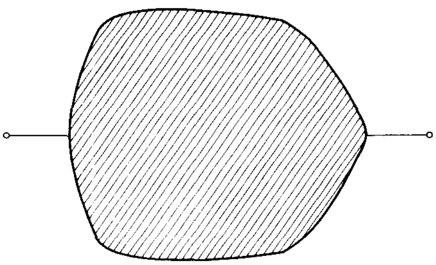

fig. 3. Monitor scope display of Collins S-Line CW waveform showing good high voltage supply regulation, which is obtainable in practice.

With under-chassis space being at a premium, three 80-μF caps (Sprague TVA 1716) were used to replace the originals. These capacitors will provide over 2½ times the former capacitance and will fit into the chassis-mounted clamps without protruding below the chassis bottom. If other types or sizes are used, be sure sufficient chassis clearance remains.

Before removing the original caps, observe the physical wiring arrangement. Place the new capacitors in the same direction as those to be removed. The original capacitors are manufactured with terminals instead of wire leads, but no difficulty should be encountered if the full-length leads of the new units are used and only the negative lead of the ground-end capacitor (C4) and positive lead of the high-voltage bus capacitor (C2) are trimmed after the connections have been made. Be sure to insulate the negative/positive interconnecting leads of the three capacitors. They are *not* at ground potential, and they pass near the capacitor mounting brackets. See **fig. 4**.

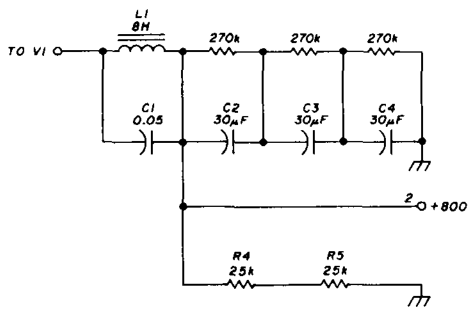

fig. 4. Modified Collins 516-F2 supply. High-voltage filter caps C2, C3, and C4 are replaced with 80-μF caps.

After installation, make certain that no short circuits exist. When the supply is reconnected to the 32-S ( ), ensure that the bias pot adjustment is correct.

Paul Pagel, N1FB

loss of torque in HAM-M rotators

Several years after I purchased my first HAM-M rotator, it started to slow down. At first I thought the slowdown was because of the extra antennas I'd added, so I increased the voltage with another transformer in series with the internal transformer. This helped for a while, but the rotator gradually slowed down again until it took five minutes to rotate 360 degrees!

After several inquiries and investigations, I found the cause. It seems that the electrolyte in the motor capacitor, C2, a 120-140 μF, 50-Vac capacitor, dries out in time. I installed a substitute capacitor, and the antennas almost took off. I removed the extra transformer. Even then, with a heavy antenna load, the 1-rpm speed was fully restored.

The motor-start capacitor is easily changed. It's conveniently located in the control box. The as-built Cornell Dubilier capacitor part number is 51172-10, priced at $2.50. A higher quality (but possibly larger in size) replacement is probably available from your local electrical or motor supply house. The capacitance should be at least 120 μF.

I've experienced this problem several times. In one case, it happened after only one year on a new HAM-M-II.

Joe Reisert, W1JR

phono plug wiring

Soldering the braid of coaxial cable, such as RG-58/U, to the shell of a phonoplug can be a messy and sometimes frustrating job. Excessive heat may be applied, resulting in the center-conductor insulation's being damaged, causing either a shorted plug or the possibility of future intermittent trouble.

A neat and virtually short-proof connection may be made using heat-shrink tubing. Instead of soldering the coaxial braid to the shell of the plug, a 2.5-cm (1-inch) length of 9.5-mm (3/8-inch) diameter heat-shrink tubing is slipped down the coax and over the braid until even with the end of the shell. The heat of a match finishes the job quickly, neatly, and permanently.

Paul Pagel, N1FB

NEW products

For literature on any of the new products, use our *Check-Off* service on page 110.

IC-280 mobile transceiver

The versatility of a microprocessor is exemplified in the introduction of the ICOM IC-280 fm mobile radio for 2 meters. Referred to as the "remotable" radio, the IC-280 actually comes assembled for immediate operation as one box. However, the same radio may be operated as separate units by removing the head and connecting the optional remote cable to each unit. You can then mount the central head in a small place where almost no other radio will fit.

"Remotability" is not the only reason to have an IC-280. The microprocessor covers all 4 MHz of the 2-meter band, plus some at both ends, in 15- or 5-kHz steps which are selected by the user or the microprocessor. In addition, there are three memory channels to store any frequency which can be programmed on the dial.

The modular 10-watt output stage has plenty of power to drive the most popular amplifiers to full output. The continuous display of frequency in transmit, receive, or memory position makes the IC-280 the easiest-to-use fm radio, and the best-performing fm radio that ICOM has designed to date. All ICOM dealers should have them in stock and on display now. See one today at your authorized ICOM dealer, or contact ICOM East, Inc., 3331 Towerwood Drive, Dallas, Texas 75234; or ICOM West, Inc., 13256 Northrup Way, Suite 3, Bellevue, Washington 98005.

multiple-output lab power supply

A new lab power supply, capable of functioning as three separate power supplies and featuring an exclusive automatic tracking circuit, has been introduced by the B&K Precision product group of Dynascan Corporation.

The new Model 1650 offers a 5-volt dc, 5-amp output, and two separate 25-volt dc outputs at 0.5 amp. The exclusive automatic tracking circuit allows the B output to "track" voltage changes of the A supply.

Designed for use with solid-state equipment where both linear and digital circuitry may be encountered, the unit's three outputs are completely isolated, offering the user full versatility to connect the outputs in series or parallel.

The valuable tracking circuit allows proportional control of the B supply when the A supply output is varied. When the TRACK mode is selected, if the B control is set at 100 per cent, the voltage level selected with the A control will appear at both the A and B outputs. If the B control is set to a 50 per cent position, the B output will be 50 per cent of the voltage at the A output. Tracking is controlled by means of a pulse-width-modulated control signal, which is coupled through an opto-isolator. This unique B&K Precision design permits complete electrical isolation of both supplies in the tracking mode.

The tracking feature of the 1650 allows it to be used as a substitute power source for breadboard and prototype circuits and other equipment. For test purposes it can provide single or simultaneously varying voltages to observe operating effects on the circuit under test. The tracking feature can also provide positive and negative voltages for operational amplifier circuits and offers a convenient means of evaluating the performance of differential amplifiers with voltage changes.

The 1650 features automatic current limiting and short-circuit protection on all ranges and outputs. IC control circuits maintain the highest reliability and stability. All power outputs use color-coded, heavy-duty, six-way binding posts.

The B&K Precision Model 1650 multiple output power supply is a money-saving alternative to separate supplies. It's available for immediate delivery at local B&K Precision distributors for $275.00. For additional information, contact B&K Precision Sales Department, 6460 West Cortland Street, Chicago, Illinois 60635.

DE-130 electronic keyer

The new DE-130 Digital Electronic Keyer, by Dynamic Electronics, Inc., is designed to provide all the features required of a high-quality keyer at a minimum cost without additional accessories. For example, an ac power

supply is included, eliminating the need for batteries or an add-on supply. The cost of an external paddle assembly is not necessary since an electronic paddle called a "touch key" is included. The electronic

More Details? CHECK—OFF Page 110

touch key works on skin resistance, and plugs into the front of the keyer. This electronic key prevents the keyer from "walking," since no mechanical motion is required to operate it. A keying relay is included to allow the DE-130 to directly key any transmitter, regardless of the type of circuit used. For code practice or actual on-the-air monitoring, an audio circuit drives a speaker mounted in the top of the enclosure.

An additional feature is a TUNE position on the volume control, which allows the transmitter to be continuously keyed for adjustment purposes. Because the circuits are digitally generated, the dot-to-dash ratio is exactly 1 to 3, eliminating the need for weight adjustments. Both dot and dash memories are included and no new characters will be accepted until the memories are cleared.

An accessory socket is provided which completely interfaces the DE-130 Keyer with the DE-131 Message Storage Unit. With the optional DE-131, the DE-130 Keyer is converted into a large, 6000-bit memory keyer with six memories which store about eighty characters in each.

The price of the DE-130 Keyer and the DE-131 Message Storage Unit is $79.95 each. The units carry a one-year warranty and may be returned for a refund during a fifteen-day trial period. For more information, write to Dynamic Electronics, Inc., P.O. Box 896, Hartselle, Alabama 35640.

DS3100 RTTY terminal

HAL Communications Corporation is proud to announce a new standard of comparison in electronic RTTY terminals — the DS3100 ASR. The new terminal features full buffering of both received and transmitted data, thus permitting preparation of transmit text while receiving, as well as storage of up to 150 lines of received text and 50 lines of text to be transmitted. The terminal also features a new screen format with twenty-four

72-character lines split to show both receive and transmit buffers, line numbering for each buffer area, on-screen status indicators to show terminal code, rate, mode, *etc.*, and a new high-contrast green P31 phosphor screen for easier viewing. The screen also uses bright/dim intensity changes to differentiate between keyboard and received data.

A total of ten "HERE IS" programmable identifier messages are available, two of which can be saved even while power is removed from the terminal. An IDENT feature allows Morse identification, regardless of the terminal's selected data code. Other features include a real-time clock, programmable answer-back (WRU), upper and lower ASCII, ASCII speeds from 110 to 9600 baud, four keyboard-operated output switches to control accessories, and a full 25-pin modem connector, for ASCII computer connections.

As in the previous DS3000 KSR V3 terminal, the new DS3100 ASR will send and receive all three data modes (ASCII, Baudot, and Morse), allows use of continuous, line, or word transmitting modes, and has synchronous idle, unshift on space, and word wrap-around. Both the electrical and mechanical features of the terminal have been completely redesigned to use a Z80 microprocessor, plug-in circuit boards, and allow easy service. A front-face legend has been added to the keytops to fully label all control functions of the terminal and simplify operation.

The keyboard and new streamlined cabinet are color coordinated in a new two-tone castle tan and chocolate brown finish. The terminal weighs 45 lbs. (20.4 kg) net (55 lbs/25 kg shipping) and can be connected for use with 120 or 240 Vac, 50 or 60 Hz power mains. The cost is $1995.00 including shipping within the United States and deliveries of the first units will start by May 1, 1979. Contact HAL Communications Corp., Box 563, Urbana, Illinois 61801 (phone 217-367-7373) for further information.

More Details? CHECK—OFF Page 110

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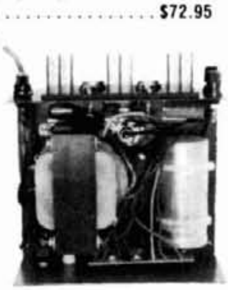
RS-7A	5	7	3¾ × 6½ × 9	8	$49.95
RS-4A	3	4	3¼ × 6 × 7½	5	$39.95

*ICS — Intermittent Communication Service (50% Duty Cycle)
If not available at your local dealer, please contact us directly.



Inside View — RS-12A

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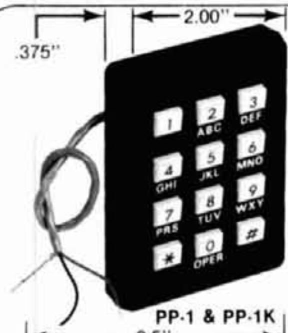
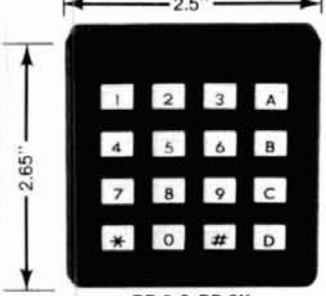
Dept. 35 P.O. Box 259
Little Neck, N.Y. 11362

No Obligation

TROUBLE FREE TOUCH TONE® ENCODER



PP-1 & PP-1K



PP-2 & PP-2K

POSITIVE TOUCH (KEYS DEPRESS) • MOBILE • HANDHELD • POSITIVE MOUNT • NO RFI • NO POTTED PARTS (SERVICEABLE) • MIL-SPEC COMPONENTS • SELF CONTAINED • XTAL CONTROLLED • LEVEL ADJUSTMENT FROM FRONT • 4.5 - 60 V.D.C. • WILL OPERATE ANY SYSTEM • LETTERING OPTIONAL

Supplied with: Instructions, schematic, template, hardware. PP-1A designed for Standard Communications Handhelds. (California orders add 6% sales tax.)

NEW PRICES PP-1 = $49.00 PP-2 = $53.00 PP1-A = $58.00
PP-1K = $55.00 PP-2K = $59.00
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Emphasis is on Quality & Reliability

TEST EQUIPMENT

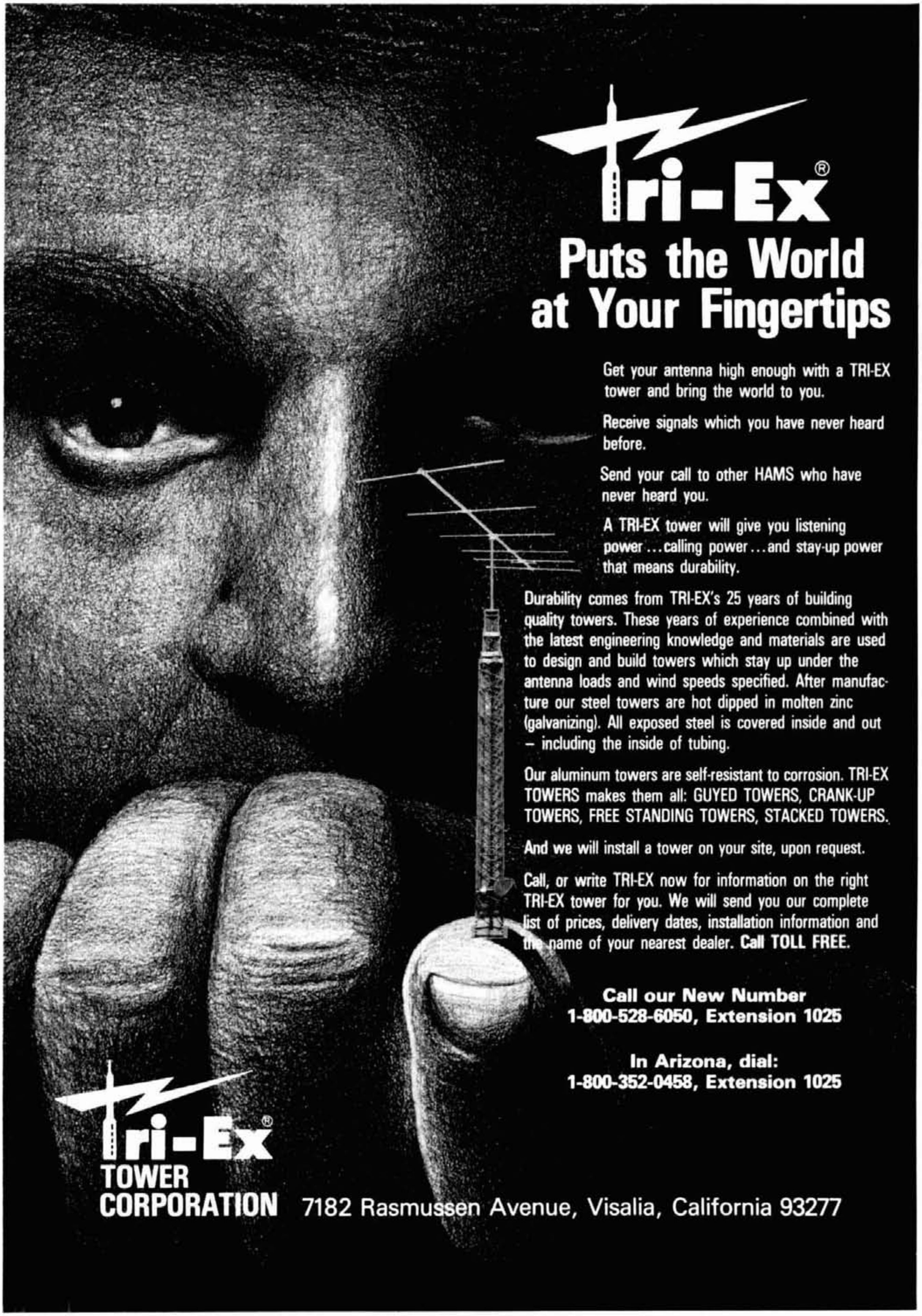
All equipment listed is operational and unconditionally guaranteed. Money back if not satisfied. Prices listed are FOB Monroe.

Boonton 190A Q-mtr 20-260MHz Q 5-1200 395
GR 1001A Stand sig gen 5kHz -50MHz calib attn 255
HP170A (USM140) 30mHz scope with reg horiz, dual trace vert plugs 475
Tek565 Dual beam 10mHz scope less plug-ins (3 series) 625
URM25 Stand Sig Gen 10kHz-50MHz calib attn 225
Weinschel 70 Prec RF attn DC-1gHz 0-60db, 1db steps 4w 395

Lists outdated. Write specifying test equipment needs. Quick response assured.

GRAY Electronics
P.O. Box 941, Monroe, Mich. 48161
Specializing in used test equipment.

More Details? CHECK — OFF Page 110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Puts the World at Your Fingertips
Get your antenna high enough with a TRI-EX tower and bring the world to you.
Receive signals which you have never heard before.
Send your call to other HAMS who have never heard you.
A TRI-EX tower will give you listening power ... calling power ... and stay-up power that means durability.
Durability comes from TRI-EX's 25 years of building quality towers. These years of experience combined with the latest engineering knowledge and materials are used to design and build towers which stay up under the antenna loads and wind speeds specified. After manufacture our steel towers are hot dipped in molten zinc (galvanizing). All exposed steel is covered inside and out – including the inside of tubing.
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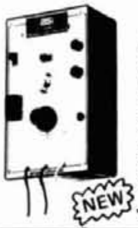
Transistor Checker



– Completely Assembled –
– Battery Operated –

The ASI Transistor Checker is capable of checking a wide range of transistor types, either "in circuit" or out of circuit. To operate, simply plug the transistor to be checked into the front panel socket, or connect it with the alligator clip test leads provided. The unit safely and automatically identifies low, medium and high-power PNP and NPN transistors. Size: 3¼" x 6¼" x 2"
"C" cell battery not included.

NEW

Trans-Check $29.95 ea.

Custom Cables & Jumpers

NEW NEW

DB 25 Series Cables

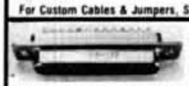
Part No.	Cable Length	Connectors	Price
DB25P-4-P	4 Ft.	2-DP25P	$15.95 ea.
DB25P-4-S	4 Ft.	1-DP25P/1-25S	$16.95 ea.
DB25S-4-S	4 ft.	2-DP25S	$17.95 ea.

Dip Jumpers

DJ14-1	1 ft.	1-14 Pin	$1.59 ea.
DJ16-1	1 ft.	1-16 Pin	1.79 ea.
DJ24-1	1 ft.	1-24 Pin	2.79 ea.
DJ14-1-14	1 ft.	2-14 Pin	2.79 ea.
DJ16-1-16	1 ft.	2-16 Pin	3.19 ea.
DJ24-1-24	1 ft.	2-24 Pin	4.95 ea.

For Custom Cables & Jumpers, See JAMECO 1979 Catalog for Pricing

CONNECTORS



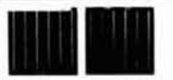
25 Pin-D Subminiature

DB25P (as pictured)	PLUG (Meets RS232)	$2.95
DB25S	SOCKET (Meets RS232)	$3.50
DB51226-1	Cable Cover for DB25P or DB25S	$1.75

PRINTED CIRCUIT EDGE-CARD

156 Spacing-Tin-Double Read-Out — Bifurcated Contacts — Fits .054 to .070 P.C. Cards

15/30	PINS (Solder Eyelet)	$1.95
18/36	PINS (Solder Eyelet)	$2.49
22/44	PINS (Solder Eyelet)	$2.95
50/100 (.100 Spacing)	PINS (Wire Wrap)	$6.95
50/100 (.125 Spacing)	PINS (Wire Wrap) R681-1	$6.95



Solar Cells 2x2cm

- 0.4 volts
- 100mA
- 41 MW

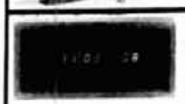
Can be added in series for higher voltage or parallel for higher current.

#SC 2x2 $1.95 ea. or 3/$5.00

the 3rd Hand

MAKES CIRCUIT ASSEMBLY A BREEZE!
Lets you work with both hands. Sturdy Aluminum Construction. **$9.95 ea.**

- **Clamp "3rd Hand" on edge of bench, table or work-board. Insert circuit board, position components.**
- **Flip circuit board to flat position for soldering and clipping.**



- Bright .300 ht. comm. cathode display
- Uses MM5314 clock chip
- Switches for hours, minutes and hold modes
- Hrs. easily viewable to 20 ft.
- Simulated walnut case
- 115 VAC operation
- 12 or 24 hr. operation
- Incl. all components, case & wall transformer
- Size: 6¾" x 3-1/8" x 1¾"

JE701

6-Digit Clock Kit $19.95

REMOTE CONTROL TRANSMITTER & RECEIVER

$19.95

INSTRUMENT/CLOCK CASE

This case is an injection molded unit that is ideal for uses such as DVM, COUNTER, or CLOCK cases. It has dimensions of 4½" in length by 4" in width by 1-9/16" in height. It comes complete with a red bezel.

PART NO: IN-CC $3.49 each

MICROPROCESSOR COMPONENTS

8080A/8080A SUPPORT DEVICES

8080A	CPU	$9.95
8212	8-Bit Input/Output	3.25
8214	Priority Interrupt Control	5.95
8216	Bi-Directional Bus Driver	3.49
8224	Clock Generator/Driver	3.95
8226	Bus Driver	3.49
8228	System Controller/Bus Driver	5.95
8238	System Controller	5.95
8251	Prog. Comm. I/O (USART)	7.95
8253	Prog. Interval Timer	14.95
8255	Prog. Periph. I/O (PPI)	9.95
8257	Prog. DMA Control	19.95
8259	Prog. Interrupt Control	19.95

6800/6800 SUPPORT DEVICES

MC6800	MPU	$14.95
MC6802CP	MPU with Clock and Ram	24.95
MC6810API	128X8 Static Ram	5.95
MC6821	Periph. Inter. Adapt (MC6820)	7.49
MC6828	Priority Interrupt Controller	12.95
MC6830L8	1024X8 Bit ROM (MC68A30-8)	14.95
MC6850	Asynchronous Comm. Adapter	7.95
MC6852	Synchronous Serial Data Adapt.	9.95
MC6860	0-600 bps Digital MODEM	12.95
MC6862	2400 bps Modulator	14.95
MC6880A	Quad 3-State Bus. Trans. (MC8T26)	2.25

MICROPROCESSOR CHIPS—MISCELLANEOUS

Z80(780C)	CPU	$19.95
Z80A(780-1)	CPU	24.95
CDP1802	CPU	19.95
2650	MPU	19.95
8035	8-Bit MPU w/clock, RAM, I/O lines	19.95
P8085	CPU	19.95
TMS9900JL	16-Bit MPU w/hardware, multiply & divide	49.95

SHIFT REGISTERS

MM500H	Dual 25 Bit Dynamic	$.50
MM503H	Dual 50 Bit Dynamic	.50
MM504H	Dual 16 Bit Static	.50
MM506H	Dual 100 Bit Static	.50
MM510H	Dual 64 Bit Accumulator	.50
MM5016H	500/512 Bit Dynamic	.89
2504T	1024 Dynamic	3.95
2518	Hex 32 Bit Static	4.95
2522	Dual 132 Bit Static	2.95
2524	512 Static	.99
2525	1024 Dynamic	2.95
2527	Dual 256 Bit Static	2.95
2528	Dual 250 Static	4.00
2529	Dual 240 Bit Static	4.00
2532	Quad 80 Bit Static	2.95
2533	1024 Static	2.95
3341	Fifo	6.95
74LS670	4X4 Register File (TriState)	2.49

UART'S

A-Y-5-1013	30K BAUD	5.95

MICROPROCESSOR MANUALS

M-Z80	User Manual	$7.50
M-CDP1802	User Manual	7.50
M-2650	User Manual	5.00

ROM'S

2513(2140)	Character Generator(upper case)	$9.95
2513(3021)	Character Generator(lower case)	9.95
2516	Character Generator	10.95
MM5230N	2048-Bit Read Only Memory	1.95

RAM'S

1101	256X1	Static	$1.49
1103	1024X1	Dynamic	.99
2101(8101)	256X4	Static	3.95
2102	1024X1	Static	1.75
21L02	1024X1	Static	1.95
2111(8111)	256X4	Static	3.95
2112	256X4	Static MOS	4.95
2114	1024X4	Static 450ns	9.95
2114L	1024X4	Static 450ns low power	10.95
2114-3	1024X4	Static 300ns	10.95
2114L-3	1024X4	Static 300ns low power	11.95
5101	256X4	Static	7.95
5280/2107	4096X1	Dynamic	4.95
7489	16X4	Static	1.75
74S200	256X1	Static Tristate	4.95
93421	256X1	Static	2.95
UPD414 (MK4027)	4K	Dynamic 16 pin	4.95
UPD416 (MK4116)	16K	Dynamic 16 pin	14.95
TMS4044-45NL	4K	Static	14.95
TMS4045	1024X4	Static	14.95
2117	16,384X1	Dynamic 350ns (house marked)	9.95
MM5262	2KX1	Dynamic	4/1.00

PROM'S

1702A	2048	FAMOS	$5.95
TMS2516 (2716)	16K*	EPROM(Intel 2716)	49.95
	*Requires single +5V power supply		
TMS2532	4KX8	EPROM	89.95
2708	8K	EPROM	10.95
2716 T.I.	16K**	EPROM	29.95
	**Requires 3 voltages, −5V, +5V, +12V		
5203	2048	FAMOS	14.95
6301-1(7611)	1024	Tristate Bipolar	3.49
6330-1(7602)	256	Open C Bipolar	2.95
82S23	32X8	Open Collector	3.95
82S115	4096	Bipolar	19.95
82S123	32X8	Tristate	3.95
74186	512	TTL Open Collector	9.95
74188	256	TTL Open Collector	3.95
74S287	1024	Static	2.95

CSC CONTINENTAL SPECIALTIES PROTO BOARDS

Proto Board 203

PB 203 **$75.00**

Model Number	L x W x H (Inches)	Price
PB-6	6.0 x 4.5 x 1.4	$15.95
PB-100	6.0 x 4.5 x 1.4	$19.95
PB-101	6.0 x 4.5 x 1.4	$22.95

Proto Board 203A

PB 203A **$124.95**

Model Number	L x W x H (Inches)	Price
PB-102	7.0 x 4.5 x 1.4	$26.95
PB-103	9.0 x 6.0 x 1.4	$44.95
PB-104	9.8 x 8.0 x 1.4	$54.95

B&K PRECISION

3½-Digit Portable DMM

- Overload Protected
- .3" high LED Display
- Battery or AC operation
- Auto Zeroing
- 1mv, 1Va, 0.1 ohm resolution
- Overange reading
- 10 meg input impedence
- DC Accuracy 1% typical

Ranges: DC Voltage – 0-1000V
AC Voltage 0-1000V
Freq. Response: 50-400 HZ
DC/AC Current: 0-100mA
Resistance: 0-10 meg ohm
Size: 6.4" x 4.4" x 2"

Model 2800 $99.95
Comes with test leads, operating manual and spare fuse

Accessories:

AC Adapter BC-28	$9.00
Rechargeable Batteries BP-26	20.00
Carrying Case LC-28	7.50

CSC CONTINENTAL SPECIALTIES CORPORATION

100 MHz 8-Digit Counter

- 20 Hz-100 MHz Range
- .6" LED Display
- Crystal-controlled timebase
- Fully Automatic
- Portable — completely self-contained
- Size — 1.75" x 7.38" x 5.63"
- Four power souces, i.e. batteries, 110 or 220V with charger, 12V with auto lighter adapter and external 7.2-10V power supply

MAX-100 $134.95

ACCESSORIES FOR MAX 100:
Mobile Charger Eliminator use power from car battery — **Model 100 — CLA $3.95**
Charger/Eliminator use 110 V AC — **Model 100 — CAI $9.95**

REGULATED POWER SUPPLY

NEW

JE200 5V-1 AMP POWER SUPPLY

- Uses LM309K
- Heat sink provided
- PC Board construction
- Provides a solid 1 amp @ 5 volts
- Can supply up to ±5V, ±9V and ±12V with JE205 Adapter
- Includes components, hardware & instructions
- Size: 3½"x5"x2"H

JE200 $14.95

JE205 ADAPTER BOARD

– Adapts to JE200 –
±5V, ±9V and ±12V

- DC/DC converter w/ +5V input
- Toriodal hi-speed switching XMFR
- Short circ. protection
- PC Brd. construction
- Piggy-back to JE200 board
- Size: 3½"x2"x9/16"h

JE205 $12.95

$10.00 Minimum Order — U.S. Funds Only
California Residents — Add 6% Sales Tax

Spec Sheets — 25¢
1979 Catalog Available—Send 41¢ stamp

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"Pennywhistle 103"

$139.95 Kit Only

The Pennywhistle 103 is capable of recording data to and from audio tape without critical speed requirements for the recorder and it is able to communicate directly with another modem and terminal for telephone "hamming" and communications. In addition, it is free of critical adjustments and is built with non-precision, readily available parts.

Data Transmission Method	Frequency-Shift Keying, full-duplex (half-duplex selectable)
Maximum Data Rate	300 Baud
Data Format	Asynchronous Serial (return to mark level required between each character)
Receive Channel Frequencies	2025 Hz for space; 2225 Hz for mark
Transmit Channel Frequencies	Switch selectable: Low (normal) – 1070 space, 1270 mark. High – 025 space, 2225 mark
Receive Sensitivity	-46 dbm accoustically coupled
Transmit Level	-15 dbm nominal. Adjustable from -6 dbm to -20 dbm
Receive Frequency Tolerance	Frequency reference automatically adjusts to allow for operation between 1800 Hz and 2400 Hz
Digital Data Interface	EIA RS-232C or 20 mA current loop (receiver is optoisolated and non-polar)
Power Requirements	120 VAC, single phase, 10 Watts
Physical	All components mount on a single 5" by 9" printed circuit board. All components included.

Requires a VOM, Audio Oscillator, Frequency Counter and/or Oscilloscope to align.

TRS-80 16K Conversion Kit

Expand your 4K TRS-80 System to 16K. Kit comes complete with:

- 8 each UPD416-1 (16K Dynamic Rams) 250NS
- Documentation for conversion

TRS-16K $99.95

COMPUTER CASSETTES

- 6 EACH 15 MINUTE HIGH QUALITY C-15 CASSETTES
- PLASTIC CASE INCLUDED 12 CASSETTE CAPACITY
- ADDITIONAL CASSETTES AVAILABLE #C-15-$2.50 ea.

CAS-6 $14.95
(Case and 6 Cassettes)

SUP 'R' MOD II

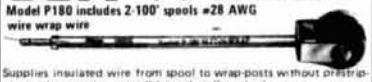
UHF Channel 33 TV Interface Unit Kit

Wide Band B/W or Color System
- Converts TV to Video Display for home computers, CCTV camera, Apple II, works with Cromeco Dazzler, SOL-20, IRS-80, Challenger, etc. MOD II is pretuned to Channel 33 (UHF).
- Includes coaxial cable and antenna transformer.

MOD II $29.95 Kit

SLIT-N-WRAP

Model P180 includes 2-100' spools #28 AWG wire wrap wire



Supplies insulated wire from spool to wrap-posts without prestripping and precutting using "daisy chain" method.

Model P180 Tool $24.50

P180A Replacement Bit	**$12.95**
W28-2 Replacement wire (3 spool pkg.)	**$2.75 ea.**

Specify color: A-green B-red C-clear D-Blue

IDEAL FOR TRS 80

CASSETTE CONTROLLER

"Plug/Jack" interface to any computer system requiring remote control of cassette functions.
The CC100 controls cassette motor functions, monitors tape location with its internal speaker and requires no power. Eliminates the plugging and unplugging of cables during computer loading operation from cassette.

#CC-100 $29.50

63-Key Unencoded Keyboard

This is a 63-key, terminal keyboard newly manufactured by a large computer manufacturer. It is unencoded with SPST keys, unattached to any kind of PC board. A very solid molded plastic 13 x 4" base suits most application. IN STOCK **$29.95/each**

Hexadecimal Unencoded Keypad

19-key pad includes 1-10 keys, ABCDEF and 2 optional keys and a shift key. **$10.95/each**

More Details? CHECK—OFF Page 110

flea market

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DEADLINE 15th of second preceding month.

SEND MATERIAL TO: Flea Market, Ham Radio, Greenville, N.H. 03048.

RM-300 RTTY MODEM, the complete TU and AFSK generator on one board. This modem featured Sept '78 HAM RADIO. Complete documentation $2.00. RM-300 Modem or RP-400 Power Supply board just $21.25. Complete RM-300 or RP-400 kit less PROM $71.25. PROM programmed with your call $7.00. WA6DNR Eclipse Communications, 5 Westwood Drive, San Rafael, CA 94901.

VERY in-ter-est-ing! Next 3 issues $1. "The Ham Trader", Sycamore, IL 60178.

MUSEUM for radio historians and collectors now open. Free admission. Old time amateur (W2AN) and commercial station exhibits, 1925 store and telegraph displays, 15,000 items. Write for details. Antique Wireless Assn., Holcomb, N. Y. 14469.

ELECTRONIC BARGAINS, CLOSEOUTS, SURPLUS! Parts, equipment, stereo, industrial, educational. Amazing values! Fascinating items unavailable in stores or catalogs anywhere. Unusual FREE catalog. ETCO-012, Box 762, Plattsburgh, N.Y. 12901. SURPLUS WANTED.

STAR-TRONICS monthly picture flyer is full of parts and pieces for the builder. Cheap. Quality. U.S. only. Star-Tronics, Box 683, McMinnville, OR 97128.

SCHEMATICS, PHOTOS, AND TECHNICAL DATA wanted by historian on merchant marine receivers made by Cardwell, Cutting and Washington, Federal, Heintz and Kaufman, Kilbourne and Clark, Signal Electric, and Wireless Egert; also on amateur receivers by Breting, Davco, Galaxy, Guthman, Lincoln, Morrow, Norden-Hauck, Patterson, Pierson, Pilot, Postal, REL, Ross, Silver. H. L. Chadbourne, 530 Midway St., La Jolla, CA 92037.

MOBILE IGNITION SHIELDING provides more range with no noise. Available most engines. Many other suppression accessories. Literature, Estes Engineering, 930 Marine Dr., Port Angeles, WA 98362.

MOTOROLA EQUIPMENT: (15) U43HHT-1100E High Band 40 Watt Motracs, no accessories, Serviceable Condition at $100.00 each. (5) Mocom 30 High Band (Low Split 136-150.8 Mc) all accessories, Serviceable Condition at $250.00 each. (10) T-1200 Remote control consoles less mics. Some have clocks, some not, Serviceable at $100.00 each. (5) Motorola single tone decoders at $35.00 each. Cal Moss, KL7HEM c/o NAVCOM AVIONICS, P.O. Box 11560, Reno, Nevada 89510 (702) 329-0019 or 786-1296.

RECONDITIONED TEST EQUIPMENT for sale. Catalog $.50. Walter, 2697 Nickel, San Pablo, Ca. 94806.

COLLINS FILTERS, $30 each, F455N-20, 40, 80, 160 with matching components, Guaranteed. Ernie Franke, WA2EWT, P.O. Box 364, Colombo, Sri Lanka (Ceylon).

MOBILE HF ANTENNA, 3.5-30 MHz inclusive, 750 watts PEP, center loaded coil, tuned from the base, eliminating coil changing or removing from mount. Less than 1.5 to 1 VSWR on all ham bands. $119.95 each — contact your local dealer or order from Anteck, Inc., Box 415, Route 1, Hansen, ID 83334. (208) 423-4100. Master Charge and VISA cards accepted. Dealer inquiries invited.

450 MHz REPEATER DUPLEXER: Decibel Products DB-4023 (3 cavities), $175. Hank Riley, 408 County Street, New Bedford, MA 02740 (617) 994-3487.

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QSL CARDS 500/$10. 400 illustrations, sample. Bowman Printing, Dept. HR, 743 Harvard, St. Louis, MO 63130.

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WANTED: Passband tuning unit from R-4A or R-4B, or complete junked receiver with tuner still intact. K1XX, *ham radio*, Greenville, NH 03048.

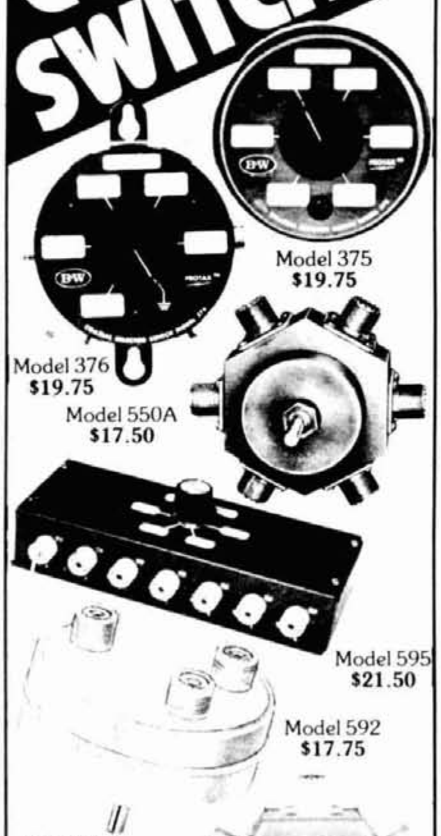
BUY-SELL-TRADE. Send $1.00 for catalog. Give name address and call letters. Complete stock of major brands new and reconditioned amateur radio equipment. Call for best deals. We buy Collins, Drake, Swan, etc. Associated Radio, 8012 Conser, Overland Park, KS 66204. (913) 381-5900.

TELETYPE EQUIPMENT for beginners and experienced operators. RTTY machines, parts, supplies. Beginner's special: Model 15 Printer and demodulator $139.00. Dozen black ribbons $6.50; case 40 rolls 11/16 perf. tape $17.50 FOB. Atlantic Surplus Sales, 3730 Nautilus Ave., Brooklyn, N. Y. 11224. Tel: (212) 372-0349.

COMPUTER FOR SALE: SWTP 6800 w/16 memory. Real-time multi-processing OS. Extra interference and memory boards, lots of software. AC-30 cassette interface included. $1200.00 takes all. May trade. KL7EWQ, Box 839, Wasilla, AK 99687.

More Details? CHECK—OFF Page 110

COAXIAL SWITCHES



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WWVB RECEIVER. The Elemek Model LXK receiver receives the 60 kHz WWVB signal and provides the following outputs: 60 kHz and 100 kHz, phase locked to WWVB carrier, and the demodulated WWVB time code. Send for brochure describing the receiver to Elemek, Inc., Dept. H, 6500 Joy Rd., East Syracuse, N. Y. 13057.

FREQUENCY ALLOCATION CHART. See how the entire radio spectrum is used. 2 kHz to 200 GHz. Send $3.00 Collins Chart Co., Box 935, Coronado, CA 92118.

WANTED: Hilltop property near Pollock Pines, California. WA6COA, 4 Ajax Place, Berkeley, CA 94708.

CALL TOLL FREE for an EZ deal. 800-247-2476/1793. Iowa call 800-362-2371. See ad elsewhere. WØEZ, Bob Smith Electronics, RFD 3, Fort Dodge, Iowa 50501.

MICROWAVE MODULES TRANSVERTER MMT 432/28S, never used, original sealed carton. $285 CDN or $250 US. Hans Peters, VE3CRU (416) 423-9446, Box 6286, Station A, Toronto, Ontario, Canada M5W 1P3.

FERRITE BEADS: w/specification and application sheet - 12/$1.00. Assorted PC pots - 10/$1.00. Miniature mica trimmers, 3-40 pF. - 5/$1.00. Postpaid. Includes latest catalog. Stamp for catalog alone. CPO Surplus, Box 189, Braintree, MA 02184.

LICENSED AMATEURS ONLY — ATV builders etc., here is your chance for brand new video equipment at dealers Cost. Retiring and closing out with such brands as Vicon, RCA and Hitachi. Write W9HVZ, check call book.

ATTENTION: UHF and Microwave Amateurs, at last a single source for your state-of-the-art Preamplifier and coaxial components. Send for free catalog. Cascade Microwave Labs, Rt. 1, Box 25, Jefferson, OR 97352.

PC BOARDS Deluxe Memory Keyer, April HR. Glass, plated, drilled, $19.95. RTC Electronics, Box 2514, Lincoln, Nebraska 68502.

WANTED: Yaesu FT-620B with manuals. Contact Charlie Perry WD5DEI, 1401 St. Johns Dr. #1 El Paso, TX 79903 (915) 565-8174.

URM-25F SIGNAL GENERATOR $200; Tektronix 212 dual trace miniscope, mint condition $650; HP-97 Calculator, like new, $550; Heathkit SB-301 receiver AM/SSB/CW filters, $175; SB-310 ham/sw reciver AM/SSB filters, $150, both receivers in excellent condition. Gene K7GK 11145 N.W. Rainmont, Portland, OR 97229 (503) 644-6156.

RTTY AFSK Modulator PC board. See Feb. 79 Ham Radio. Drilled $5.00 F. E. Hinkle, 12412 Mossy Bark, Austin, TX 78750.

EZ deals are the best! Try me and see for Yaesu, Drake, KLM, Swan, Cushcraft, DenTron, VHF Eng., ICOM, CDE, Hustler, Wilson and more. Call, see or write WØEZ, Bob Smith Electronics, RFD 3, Hwy 169 & 7, Fort Dodge, IA 50501. (515) 576-3886.

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MINT SB200 — hardly used 80-10M. Dust cover. $425 ppd. F. E. Hinkle, K5PA, 12412 Mossy Bark, Austin, TX 78750.

WANTED: Motorola KXN 1024 and KXN 1052 channel elements. WA6COA, 4 Ajax Place, Berkeley, CA 94708.

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Collins 312B5, Vfo Console, exc	$525
Collins 75S3C, rnd, exc	$1250
Collins 180S1, antenna tuner	$250
Collins 30L1, vy gd	$595
Collins 516F2, power supply	$185
Collins R-390A rcvrs, overhauled exc cond	call for quote
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Hammarlund SP-600JX, rcvr	$395
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Test Gear

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Measurements Model 80, 2-400 MHz	$295
Measurements Model 65B signal generator	$295
HP-202H 54-216 MHz AM/FM sig gen	$695
HP-608D 10-420 MHz sig gen	$550
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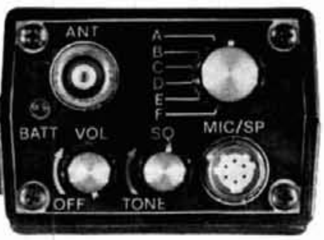

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WANTED: GE MASTR ER-41C receiver, ET-58 transmitter. Fred Findling, P.O. Box 125, Littleton, N. H. 03561.

RTTY — Bandpass active filter 2125/2295 Hz. Kit $11.95, W/T $16.95 ppd. SASE for info. Nat Stinnette Electronics, Tavares, FL 32778.

REPEATERS — Factory Built VHF Engineering, 4 uhf, 1 with P.L. Two in unopened boxes. 1 220 MHz, 2K Takes all. WA6COA, 4 Ajax, Berkeley, CA 94708.

THE MEASUREMENT SHOP has used/reconditioned test equipment at sensible prices; catalog. 2 West 22nd St., Baltimore, MD 21218.

WANTED — Instructor Ham Radio. N.Y.S. Co-ed Children's Sleep-Away Camp. Write: Camp Kinder Ring, 45 East 33rd Street, New York, N.Y. 10016.

MANUALS for most ham gear 1937/1970. Send 25¢ for "Manual Catalog." HI, Inc., Box H864, Council Bluffs, Iowa 51502.

100 PLUS BB403 NICAD BATTERIES unused, individ. packed with vent cap 4 Amp hr. 1.2V per cell requires electrolite. Also 126 New SO 239 Conn. Will sell either separate or in quan. Open to offers. E. L. Petrick, P.O. Box 1255, Oak Lawn, IL 60454.

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WANTED — Radio transcription discs. Any size or speed. Larry, W7FIZ. Box 724, Redmond, Washington 98052.

TEST EQUIPMENT CATALOG listing used Tektronix, HP and GR equipment at bargain prices. PTI, Box 8699, White Bear Lake, MN 55110. Price $1.00 refundable with first order.

Coming Events

MASSACHUSETTS: NoBARC Hamfest July 21, 22 at Cummington Fairgrounds. Tech talks, demonstrations and dealers. Flea market, $1.00; advance registration $3.00 single, with spouse $5.00 to Tom Hamilton, WA1VPX, 206 California Avenue, Pittsfield, Massachusetts 01201 or $4.00/$6.00 at gate. Mobile talk-in on 146.31/91. Gates open at 5:00 PM on Friday for free camping.

ARKANSAS: The Central Arkansas Radio Emergency Net's Second Annual Ham-A-Rama, August 4 and 5, Arkansas State Fairgrounds, Little Rock. Talk-in on 146.34/94. Contact: Morris Middleton, AD5M, 19 Elmherst Dr., Little Rock, AK 72209 (501) 568-0938.

CONNECTICUT: The CQ Radio Club of Torrington's Flea Market and Hamfest on June 17, 9 AM to 5 PM, Torrington Fish and Game Association Grounds, Weed Rd., just off Route 4, between Torrington and Goshen. Refreshments available, door prizes, parking and table space. Children's activities. Admission: $1.00 per person includes vehicle. Spouses and children free. Talk-in on 52 direct and 147.84 TR - 147.24 Rec. Call K1BCI for directions. For details: Ed Josefow, W1JSU, (203) 482-1837.

OHIO: Official ARRL 5th Annual Hall of Fame Hamfest. Sunday, July 15, Stark County Fairgrounds, Canton, Ohio. Mobile check-in on 19-79 and 52-52. $2.50 Advanced, $3.00 at gate. Contact WA8SHP 10877 Hazelview Ave., Alliance, OH 44601.

OHIO: Sandusky Valley Radio Club's Hamfest, Sunday, June 10, Sandusky County Fairgrounds, Fremont. For info: John Dickey, W8CDR, Sec., 545 N. Jackson St., Fremont, OH 43420.

OHIO: The Tusco Radio Club and the Canton Amateur Radio Club present the 5th annual Hall of Fame ARRL Hamfest, Sunday, July 15, Stark County Fairgrounds, Canton. For info: Hall of Fame Hamfest, 10877 Hazelview Ave., Alliance OH 44601.

PENNSYLVANIA: Milton ARC Hamfest, June 3, 1979 at the Allenwood Firemen's Fairgrounds. U.S. Route 15, four miles north of Interstate 80. Doors open 8 AM $2.50 advanced, $3 gate; children and spouses FREE. Flea market, auction, contests, portable/mobile FM clinic, prizes, food & beverages. Talk-in on 146.37/.97; 146.34/.94 repeaters and 146.52 simplex. Details from Kenneth Hering, WA3IJU, R.D.#1, Box 381, Allenwood, Pennsylvania 17810; Telephone (717) 538-9168.

PENNSYLVANIA: The Tri-Club Hamfest (W3OK, WA3GYE and W3OI) July 15, 8 AM to 4 PM, Allentown Police Academy pistol range, Lehigh Parkway South, Allentown. Admission: $2.00 lookers; $4.00 sellers. Talk-in on .34/.94 and .52.

More Details? CHECK—OFF Page 110

More Details? CHECK—OFF Page 110

ILLINOIS: *The Quad County Radio Club's 22nd Annual* Breakfast Club Picnic and Hamfest, Saturday and Sunday, July 14 and 15, Terry Park, Palmyra. Entertainment, discussions, meetings. Open air flea market. Prize drawing 3:00 PM Sunday. For info: Quad County Radio Club, *P.O. Box 81, Chatham, IL 62629.*

INDIANA: Hamfest, Sunday, July 8, Marion County Fairgrounds, Indianapolis eastside, intersection of I-74 and I-465. Commercial exhibitor area, indoor/outdoor *flea market. Addition indoor facilities.* Tech forums, ladies' activities, refreshments available. Overnight parking and hookup for RV's on grounds. Talk-in on 16-76, 28-88, 10-70. Additional info: Indianapolis Hamfest Association, Box 1002, Indianapolis, IN 46206.

MARYLAND: The Frederick Amateur Radio Club's Second Annual Central Maryland Hamfest, June 17, Frederick Fairgrounds, Frederick. For info: Frederick Amateur Radio Club, Inc., Box 1260, Frederick, MD 21701.

MICHIGAN: Chelsea Swap 'n Shop, Sunday, June 3, 1979 at the Chelsea Fairgrounds. Gates open to sellers, 5 AM; to public, between 8 AM and 3 PM. Admission $1.50 advance, $2 at gate; children under 12 and non-ham spouses admitted free. Proceeds to benefit Dexter High School Radio Club and Chelsea Communications Club. Talk-in on 146.37-.97.

31ST ANNUAL U.P. HAMFEST will be held on July 28 & 29, 1979. Add your name to our mailing list. Send QSL to Sawyer Radio Ass., P.O. Box 73, Gwinn, MI 49843.

MISSOURI: Indian Foothills A.R.C. Hamfest, July 22, 1979, Saline County Fairgrounds, Marshall, Missouri. *Gates open 8 AM. Tickets $2 each, or 3 for $5, advance;* $2.50 each, at door. Flea Market (no charge for tables), prizes, displays, campground (no utilities), food, and activities for the entire family. Talk-in on 146.52 simplex; 146.28/.88 and 147.84/.24 repeaters. Information and *tickets from Norman Gibbins, WBØSCI, 692 North Ted,* Marshall, Missouri 65340.

MONTANA: The International Glacier-Waterton Hamfest, July 21 and 22, Three Forks Campground, 10 miles east of Essex on U.S. Highway 2. Registration: 9 AM. Talk-in on 52 and 34/94. For info: Glacier-Waterton Hamfest, P.O. Box 2225, Missoula, MT 59806.

NEW JERSEY: Shore Points Amateur Radio Club's 2nd Annual Hamfest and Electronic Flea Market, Sunday, June 10, 8 AM to 4 PM, rain or shine, Stockton State College Campus, Pomona. Just 12 miles west of Atlantic City Boardwalk and Casinos. Registration: $2.00 per person, under 12 free. Tailgating: $2.00 per car space. Indoors: $5.00 per space. For info: Monte Tremont, WB2EYF, P.O. Box 142, Absecon, N.J. 08201 (609) 266-2678.

NORTH CAROLINA: The CARY Amateur Radio Club's Seventh Annual Mid-Summer Swapfest, Saturday, July *21, Lion's Club Shelter, Cary. Buy, sell, trade electronic* equipment. No commission. 9 AM - 3 PM. Talk-in on 146.28/.88, 147.57/.25, 146.52/.52. For info: CARY ARC, P.O. Box 53, Cary, N.C. 27511.

OHIO: *The Northern* Ohio Amateur Radio Society's Second Annual NOARSFEST, Saturday, July 7, Lorain County Fairgrounds, Rt. 18, Wellington. Over 100 prizes including new DenTron HF-200 Xcvr, Ten-Tec 509, DenTron GLA 1000, Wilson Mark II and more. Indoor hall for *dealers ($4.00 Adv.) Flea market space* $1.00 each. Gates open for sellers/dealers 6:00 AM. Public 7:00 AM to 5:00 PM. Admission: $1.50 advance, $2.00 gate. Under 12 free. Prize drawing tickets $1.00. Free camping outside gates Friday night — no hookups. Advance registration, *tickets,* info: NOARSFEST, P.O. Box 354, Lorain, OH 44052.

OHIO: Second Annual Salem Area Hamfest, 9 AM - 3 PM Sunday, August 5, Kent State Salem campus, Salem. Advance tickets, $1.50; $2.00 at door. Inside tables, $5.00, space for yours, $2.00. Flea Market space, $1.00. Air conditioning, wheelchair ramp, free parking, refreshments, prizes. Grand prize: Atlas RX-110, TX-110, PS-110. Check in 146.52 simplex. Details: Harry Milhoan, WA8FBS, 1128 West State, Salem, OH 44460.

RADIO EXPO '79, September 15 and 16, Lake County Fairgrounds, Routes 120 and 45, Grays Lake, Illinois. Manufacturers' displays, flea market, seminars, ladies' programs. Advance tickets $2.00. Write EXPO, P.O. Box *305, Maywood, IL 60153. Exhibitors inquiries: EXPO* Hotline (312) 345-2525.

WISCONSIN: The South Milwaukee Amateur Radio Club's annual SWAPFEST '79, Saturday, July 14, Ameri*can Legion Post #434,* 9327 S. Shepard Avenue, Oak Creek, 7:00 AM to 5:00 PM. Admission: $2.00 includes happy-hour with free beverages. $100 first prize, $50 second prize and others. Parking, picnicking, overnite camping. Talk-in on 146.94 MHz FM. For details including *map: South Milwaukee Amateur Radio Club,* Robert Kastelic, WB9TIK, Sec., P.O. Box 102, South Milwaukee, WI 53172.

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PENNSYLVANIA: Harrisburg RAC Annual Firecracker Hamfest, Wednesday, July 4. Shellsville VFW Picnic grounds, I-81 North, Exit #27 or Racetrack Exit #28. Admission: $3.00. Free tailgating. Tables available in pavillion. W3UU Talk-in on 52-52 or local repeaters.

TENNESSEE: The Oak Ridge Amateur Radio Club will hold the Oak Ridge Amateur Radio Convention and Hamfest '79 in Oak Ridge this year instead of Crossville. The date is July 14 and 15 at the Oak Ridge Civic Center. 10,000 square feet of air-conditioned area available for commercial and flea market exhibits. The FCC will be in Oak Ridge Saturday, July 14 to give exams. Exams will start promptly at 8 AM in the High School cafeteria across the street from the Civic Center. Anyone wanting to take the exam must have Form 610 completed and have your present license with you. Other activities planned for ladies and kids include movies, tour of Oak Ridge's world famous American Museum of Science and Energy. Or you might want to take advantage of the heated indoor pool or picnic and playgrounds at the Civic Center. By moving the Oak Ridge Amateur Radio Convention and Hamfest from Crossville to Oak Ridge, the Oak Ridge Club plans to expand in every way to make for a bigger and better Hamfest for all. The talk-in station will operate on 146.88. Other frequencies are 147.72 and 146.82. Local talk-in on 146.52. Camping facilities complete with everything are available just 20 minutes from the Civic Center. Motels and restaurants are conveniently located also. Commercial and flea market exhibitors are urged to make early arrangements to attend by contacting Charles Byrge, WB4OBE, P.O. Box 291, Oak Ridge, TN 37830. A welcome center operated by the Chamber of Commerce will be located at the Civic Center to provide visitors with full information on what to see while in Oak Ridge. The week of July 9-16 will be proclaimed Amateur Radio Week in Oak Ridge by the mayor. An Amateur Radio station will be on exhibit at the museum. Museum admission is free. Admission to Hamfest $1.00.

TEXAS: The TEX-LA Hamfest, July 7 and 8, Red Carpet Inn, Beaumont; Motel reservations close June 22. Flea market parking $3.00. Preregistration $4.00 ($6.00 at door). Flea market parking $4.00 at door. Banquet/dance Saturday night. Lots of FREE refreshments both days. Technical sessions, ladies' programs, displays of Ham gear.

ISLE OF MAN MILLENIUM DXPEDITION: The Liverpool and District Amateur Radio Society are mounting a DXpedition to the Isle of Man during the period of the special "GT" prefix to celebrate 1,000 years of the Island's Parliament, Tynwald. The club's callsign, GT3AHD/Portable, will be used during this period (2300 GMT, 29 June to 2300 GMT, 8 July, 1979) and operation will be on all HF bands (160-10 meters) and 2M/70 cm. Suggested frequencies, plus or minus QRM, are: CW 1.820-835, 3.505, 7.005, 14.080, 21.080, 28.080. SSB 1.820-835, 3.695/780, 7.092, 14.195/275, 21.245/275, 28.495/550, 144.290, 432.290. Club members operating include G3's XSN, YBH, G4's AHS, AMX, CVZ, EST, FPB, GEB, GHS, GHT, GKF, HGT, HSF, G8's AVJ, CFM, FHD, NNX, NRD. Participation in the R.S.G.B. VHF national field day is intended. QSLing will be 100% either via bureau to G3AHD or direct (including the appropriate IRC's) to: G. H. Cohen, G4GHS, 41, South Station Road, Gateacre, Liverpool, Merseyside L25 3QE, England.

AUSTRALIA: New Ten-Meter Sovereign Hill Award commemorates the foundation of the Sovereign Hill Historical Park in Ballarat, Victoria (VK3), scene of the great gold rush of the 1850s. Begins Saturday, May 12, 1979 with commencement of Sovereign Hill Amateur Radio Station on the grounds. Contact five of the award "Charter" stations on ten meters. One of these must be a local station, designated by "S" following the charter number. All amateurs winning the award will be given an "A" number which may be passed on to other Amateurs anywhere in the world toward their awards. Cost of the award is $2 U.S., or equivalent, which includes airmailing to the recipient. Frequencies and times: Transmissions by the Sovereign Hill station on public holidays and selected weekends, at a frequency of 28.530 MHz plus or minus QRM. Any contact with this station will count as two "S" contacts toward the award. All other "S" and "A" contacts count as one. In addition there will be a contact every Saturday (U.S.) at 0000 GMT in conjunction with the Welcome Stranger Ten-Ten Net, counting as one contact. Other contacts may be anywhere in the ten-meter band. For information write Leo McPherson, VK3NIQ, P.O. Box 247, Ballarat East 3350, Victoria, Australia.

KENTUCKY: The 7.228 MHz evening net will hold its annual get together this year at Barren River Lake State Resort Park, Lucas, Kentucky, June 8, 9, and 10, 1979. Dinners at 6 PM Friday and Saturday in the reserved banquet room at the Barren River Lodge. Door prizes, drawings, presentation of the Elephant Key, trips to historical sites, and fun for the entire family. Talk-in on 7.228 MHz, 146.34/.94 (Glasgow repeater) and 147.93/.33 (Bowling Green repeater). Trailers and campers invited. More information from Jack Tinney, WØJJS, 3404 South Leslie, Independence, Missouri 64055.

More Details? CHECK — OFF Page 110

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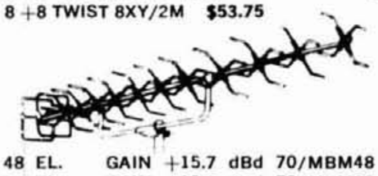

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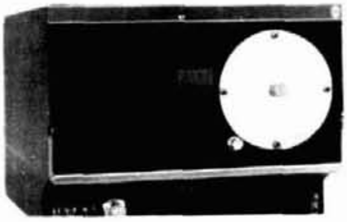

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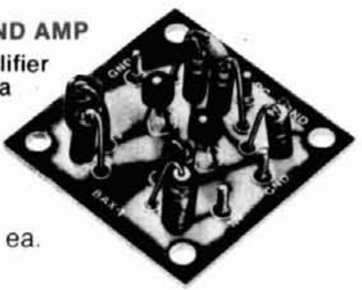

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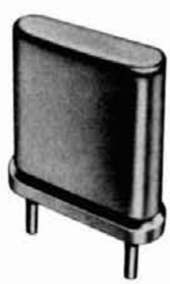

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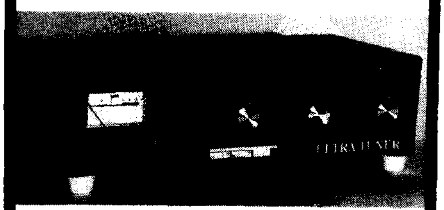

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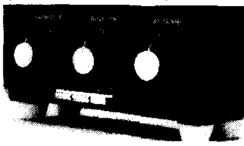

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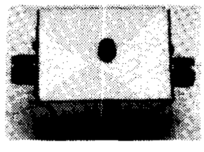

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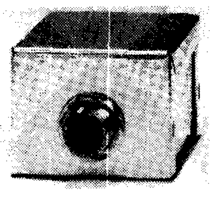

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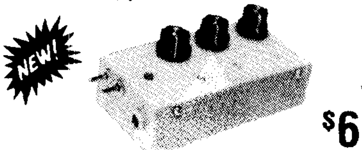

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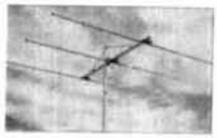

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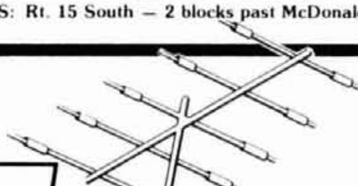

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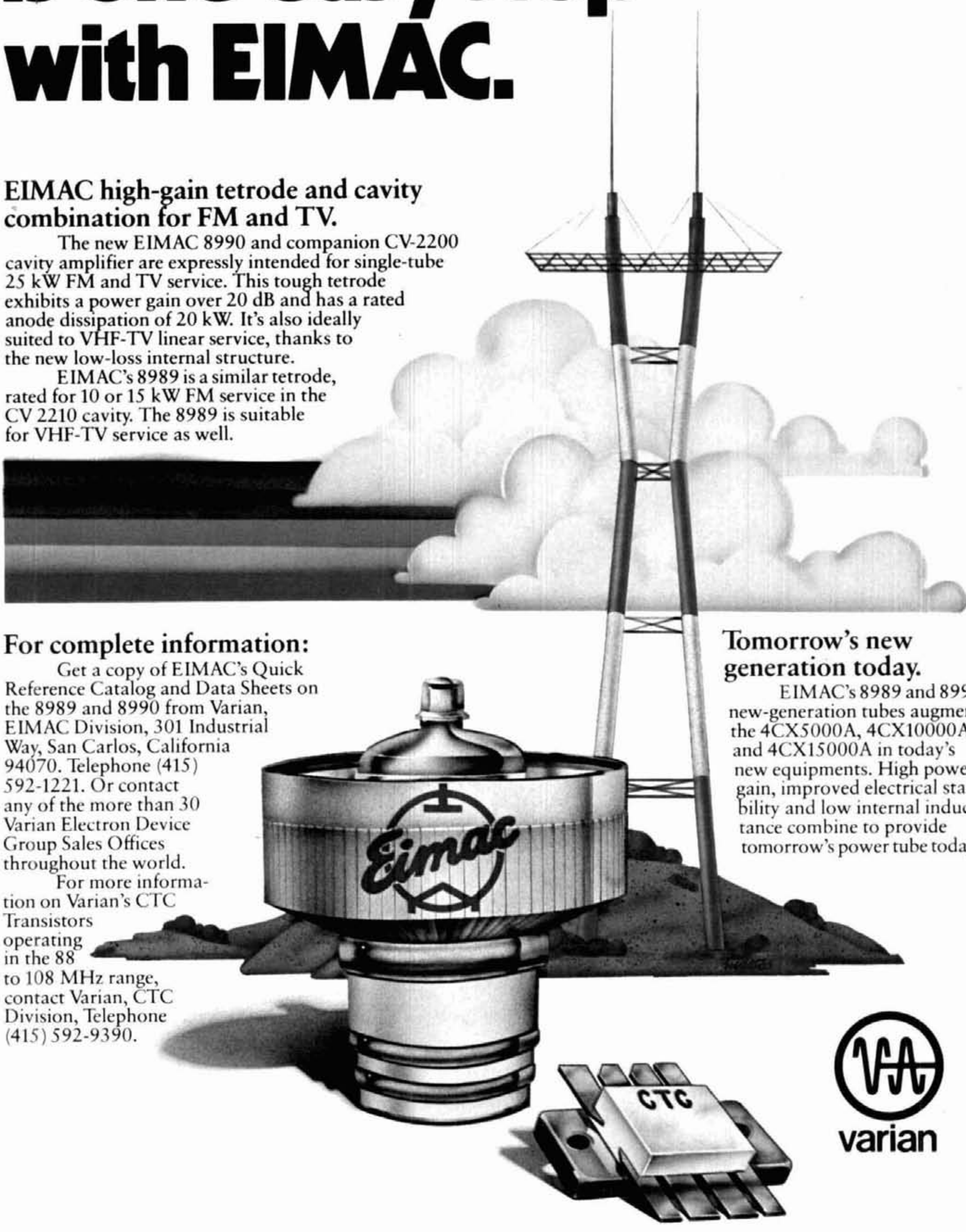

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